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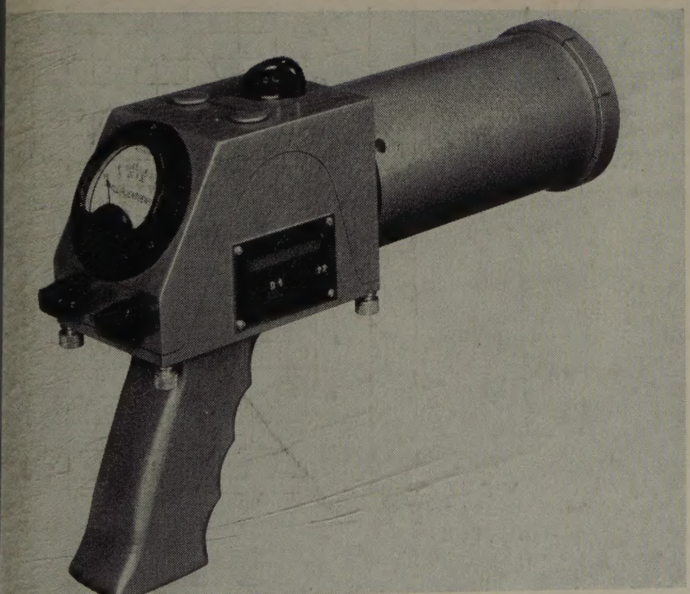

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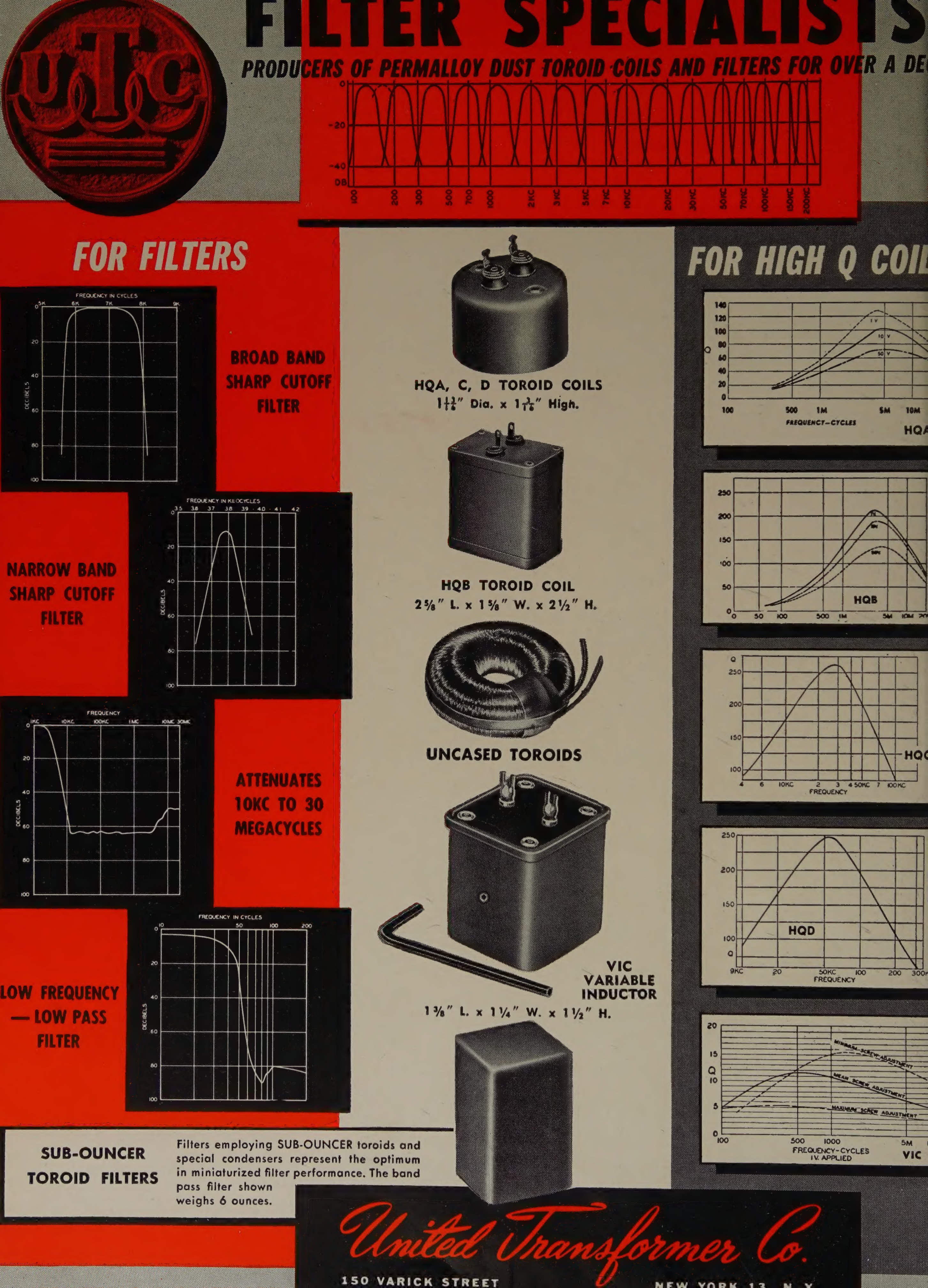
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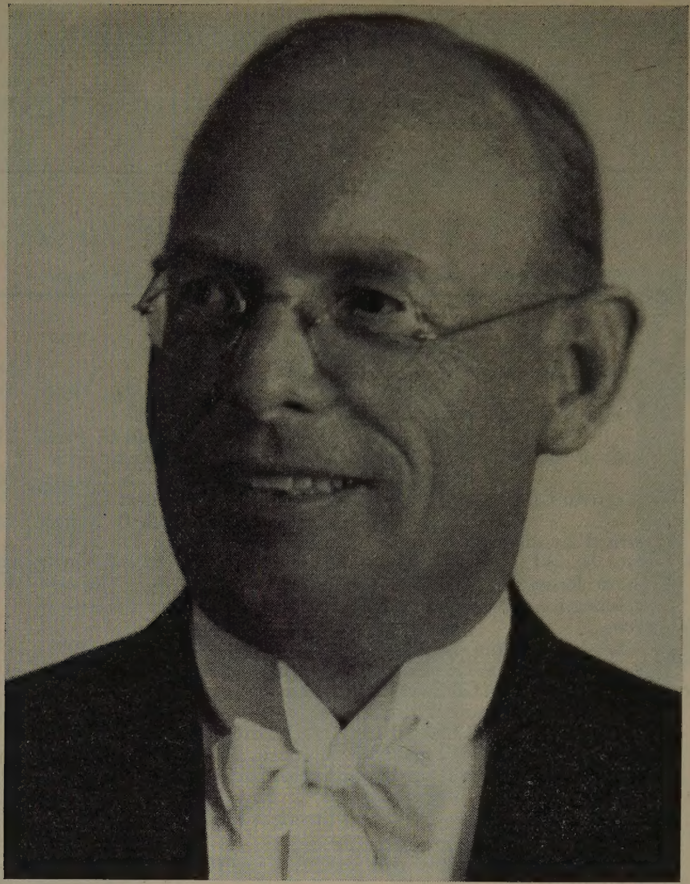
PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION



Jorgen Rybner

VICE-PRESIDENT, 1951

Jorgen Rybner was born at Frederiksberg, Denmark, on November 26, 1902. He was graduated as an electrical engineer, specializing in telecommunications, from the Royal Technical University of Denmark in 1926, and worked as a private assistant to Professor P. O. Pedersen at the university for 18 months, carrying out most of the computation work for Professor Pedersen's famous book, "The Propagation of Radio Waves."

During 1927–1928, Mr. Rybner spent a year in the United States, studying radio techniques with Professor Morecroft at Columbia University. He did part-time work in the propagation department of the Bell Laboratories, Inc. He also worked in the laboratory of the General Electric Company at Schenectady, N. Y.

He was associated with the Geodetic Institute of Denmark until 1939, serving as chief of the technical department, and he took part in the longitude determination of the Baltic Geodetic Commission 1929, carrying out extensive studies of the theory of seismographs, on which he published several papers.

In 1939, the chair of telecommunications at the Royal Technical University of Denmark held by Professor Pedersen since 1907, was doubled, and Mr. Rybner obtained the new professorship.

He has published, besides a number of scientific papers, a series of textbooks in Danish, covering telegraph and telephone techniques, including switching and application of probability, radio techniques, electronic amplifiers, network and transmission lines, and advanced books on filter theory and on the general theory of circle diagrams. Publications with English and Danish text include his "Nomograms of Complex Hyperbolic Functions," 1947, as well as "Table for Use in the Addition of Complex Numbers," 1949.

In 1948–1949, Professor Rybner paid another visit to the United States to study communication theory and advanced network theory at Massachusetts Institute of Technology, Boston, Mass., during the fall term. He also spent three weeks at Bell Telephone Laboratories, Murray Hill, N. J., and Holmdel, N. Y., with visits to the Watson Scientific Computing Laboratory, New York, N. Y., and Cornell University, Ithaca, N. Y.

Since 1933, he has been a teacher of telecommunications for signal officers at the Army Officers' School.

He was elected a member of the Danish Academy of Technical Sciences in 1937, and was president of the Danish National Committee of the International Scientific Radio Union (URSI) in 1948. He has been a chairman of the board of directors of the Radio Receiver Laboratory of the Academy since 1944. He was decorated with the Knight's order of Danebrog in 1948. Mr. Rybner has been a member of the IRE since 1926.

The dimensions—or certain other characteristics—of physical things may have chosen values. But if such values are selected from an orderly and adequate list of "preferred numbers" a certain degree of simplification, interchangeability, uniformity, convenience, and economy will result.

The writer of the following guest editorial is particularly well qualified to discuss this subject analytically. He has served as the IRE representative on the ASA Sectional Committee Z17 on Preferred Numbers, and also on the ASA Sectional Committee C67 on Standardization of Voltages—Preferred Voltages—100 Volts and Under. He is a Past President of The Institute of Radio Engineers, and an engineer of long, diverse, and constructive experience.—*The Editor.*

Preferred Numbers

ARTHUR VAN DYCK

"Preferred Numbers" is the title given to a list of numbers standardized by the American Standards Association. The intention is that designers, and others who have discretion in design choices, will use the standard numbers, wherever practicable to do so, for dimensions, weights, ratings, and other requirements for which numerical values must be specified. Of course, there may be cases where it is impracticable to use the standard numbers, for technical or economic reasons. In such cases they should not be attempted. However, by adopting standard values whenever practicable, it will be found, as time goes on, that the impracticable cases will become rarer, and the benefits will increase.

As Preferred Numbers and their advantages have been described at length elsewhere,[1–3] it will suffice here to say that use of the system by even one manufacturer will improve his efficiency of operation, and that wider use by a whole industry, or many industries, brings benefits in geometric proportion to the extent of use. The subject has had active study for about thirty years, but in spite of this lengthy period, has not made great progress toward adoption by American industry.

The Preferred Numbers activity began just after World War I, in a period when high ideals inspired the thinking elements of mankind. Even the ultimate in human relations, international amity, seemed within reach. In the thirty years that have passed—enough to have brought impressive results, it would seem—progress has been disappointingly small. The League of Nations, expected to bring peace to the world, instead saw a second World War, and its successor, the United Nations is dallying with a third. Similarly, Preferred Numbers has not made the progress which its obvious advantages make it reasonable to expect.

It seems to me that the factors which have determined the amount of progress toward world peace, or lack of it, are much the same ones which have limited the degree of accomplishment in standardization, and that it is easier to see some of these factors in the broad field of international relations than in the narrow field of product design, in spite of the vastly greater scope of the former. The greatest obstacle to international peace agreement is the unwillingness of nations to limit "national sovereignty" to even the slightest extent. Each nation insists upon full rights to have and to do whatever it wants. Little attention is paid to the long term benefits which would accrue to each and to all by the establishment of unselfish, co-operative practices. Few realize that large, permanent benefits will result from small temporary sacrifices, and that such benefits are obtainable only in that way. Most want to eat the cake and have it, too. Consequently, petty bickering continues and so do the resulting wars.

The same attitude has existed in the design field in relation to Preferred Numbers. It is clear that the same numbers must be used by everyone. If each company, or industry, has its own private list of numbers, the benefits will be greatly limited. Nevertheless, various organizations have seen fit not to adopt the ASA Preferred Numbers, but instead have set up their own lists of numbers. They have done this because their own practices *of the moment* seemed to be accommodated better by lists different from the ASA Standard. I emphasize "of the moment" because that is the heart of the matter, the seat of the trouble, the crux of the situation. For example, some years ago, the Radio-Television Manufacturers Association found it desirable to standardize the values of resistors. The ASA Preferred Numbers Standard was considered, but judged not to suit the manufacturing conditions and the buying practices of the resistor field at the moment, whereas a special series of numbers suited better. The special series was adopted and, since it was an official RTMA list, it has been utilized by later RTMA committees for other applications than resistors, although adopted originally because of seeming advantages for resistors. Ironically, the original advantages have largely disappeared through changes in resistor manufacturing conditions. But the irregular standard remains, and in fact is now being proposed for universal application in the radio-electronic industry. We in the industry, of course, feel that radio-electronics is the outstanding element in the American industrial scene, but it is at least a wee bit unfortunate that electronics, just now entering into increasingly important and intimate relationships with all other elements of industry, exercises its "sovereignty," sets its own Standard, and exchanges the greater good for the appeal of the immediate.

Another point which seems to be overlooked by most standardization committees is that a standard must have much of the aspect of an ideal if it is to be worth while, and that if it cannot be used immediately for some practical reason, *it need not be used immediately*. As time goes on, with effort to use the standard wherever conveniently possible, the first practical objections lessen, and the standard becomes more and more universally usable. After five, ten, or twenty years, the standard is wholly acceptable and its benefits are had. A good standard is a goal to shoot at, not a ratification of old or existing practice. On the other hand, if standardization work operates on the basis of setting up standards with primary attention to the operating practices of the moment—so that everything every manufacturer is doing will be "standard" immediately—little is accomplished toward the proper objective of maximum simplification. It is not simplification of current practice to standardize all of current practice. The extreme to which overlooking this basic principle naturally leads, is exemplified by the proposals now active in the work to establish a Standard for Preferred Voltages, 100 volts and under. The proposals call for standardizing more than ninety values between zero and one hundred! The original committee "ideal" list called for twenty-three values, which is big enough, but as each industry element added the values it is using currently, the list grew until nearly every number up to one hundred is proposed as standard. That surely is the height of something or other, and one might ask "How absurd can we get?" The answer to that one is easy, however. It is found in the list of current television picture tube sizes!

In the long, difficult period ahead of us, when our industrial economy will have to bear strains greater than ever before experienced, the benefits to be obtained by standardization and simplification will be needed. Preferred Numbers can help much, if they are attempted with understanding, unselfishness, and co-operation. It must be realized that there is only one list of Preferred Numbers and that one must be the national standard. Individual industry lists are not Preferred Numbers and should not be so called—they are private numbers affording a limited amount of benefit or temporary advantage, but little of the large scale benefits which will result from national, widespread, inter-industry use of one list. The production capability of this country, great as it is, is finite, not infinite, and the tough task before us demands maximum efficiency to assure success.

1 ASA American Standard Z 17.1—1936.
2 John Gaillard, "Preferred numbers—an international tool for standardizers," *Standardization*, vol. 20, pp. 296–297, 299; November, 1949.
3 Arthur Van Dyck, "Preferred numbers," Proc. I.R.E., vol. 24, pp. 159–179; February, 1936.

Television Broadcasting in the United States, 1927-1950*

DONALD G. FINK†, FELLOW, IRE

A study of the wishes of representative portions of the membership of The Institute of Radio Engineers has led to a plan to present to the readers of these PROCEEDINGS a series of tutorial papers on a wide variety of topics of both present and historical interest. These papers are to be educational in nature, of exceptional clarity, and prepared in each case by an authority in the corresponding field. It is believed that they will as well be appealing and interesting to others than experts in that field.

The procurement of these tutorial papers, and their individual recommended approval for publication, are carried out by the Subcommittee on Tutorial Papers (under the Chairmanship of Professor Ernst Weber) of the IRE Committee on Education (under the Chairmanship of Professor Herbert J. Reich). The first tutorial paper of what is hoped to be an extensive series is here presented.—*The Editor.*

TELEVISION broadcasting in America began in 1927, when the Federal Radio Commission issued the first television license to Charles F. Jenkins authorizing broadcast transmissions from a station in the suburbs of Washington, D. C. Prior to that time, development of television techniques was not open to public participation. V. K. Zworykin applied for a patent on his iconoscope, which may fairly be called the cornerstone of modern television, in 1925. In 1923, Jenkins in America and John Baird in England had demonstrated the transmission of crude images over wires. Early in 1927 the Bell Telephone Laboratories demonstrated a low-definition picture over wire circuits, between New York and Washington. But the concept of providing broadcast emissions, available to experimenters not otherwise connected with the transmitting organization, did not gain wide currency in America until 1929. In that year some 22 stations were authorized by the Federal Radio Commission to broad visual images.

The earliest stations had wide latitude in choice of frequency, almost any frequency above 1,500 kc being permitted if no interference was caused to other services. But this latitude was soon withdrawn, as the short-wave region became crowded with other, more vital services. In 1929, emissions were limited to a bandwidth of 100 kc, within the regions of 2.0–2.3 Mc and 2.75–2.95 Mc. The powers employed varied from 100 w to 20 kw, the majority of stations operating at 5 kw.

The quality of the early images were primitive, judged by any standard. The pictures were commonly transmitted at a rate of 20 per second. At this rate the number of picture elements capable of being transmitted by double sidebands in a 100-kc band is limited to 5,000. Equal resolution in vertical and horizontal dimensions was achieved within the band limits by employing a square image of about 70 lines, but the preferred figure was 60 lines.

* Decimal classification: R583. Original manuscript received by the Institute, November 15, 1950.

† *Electronics* Magazine, New York, N. Y.

Many of the major American television stations of the present day can trace their origin to this early period. The National Broadcasting Company's station in New York was first licensed as W2XBS in July, 1928, and has since evolved from the 2,000 to 2,100-kc band to the 66 to 72-Mc band, from 60-line pictures to 525-line pictures. In 1942 the call letters W2XBS were withdrawn in favor of the "commercial" call letters WNBT. Similarly, the Columbia Broadcasting System station in New York, now WCBS-TV, started in July, 1931, as W2XAX. This station operated with 60-line pictures, 20 per second, for a total of 2,500 hours in the period ending February, 1933. The General Electric station in Schenectady operated on similar standards, with 20-kw power from 1929 to 1932.

In 1931 it was evident that progress could not be made on the restricted channels of the 2-Mc band, and the trend toward higher frequencies began. One of the earliest to apply for permission to use frequencies above 40 Mc was the Don Lee Broadcasting System in Los Angeles, Calif. In December, 1931, the license of station W6XAO was granted to this organization, authorizing the use of the bands 43–46, 48.5–50.3, and 60–80 Mc. In 1941 this station became a commercial station with the call letters KTSL, operating on 54-60-Mc with a regular public program service.

Permission to use the 43-80-Mc bands was granted to several other stations, including NBC's W2XF and W2XBT in New York, Jenkin's W3XC in Wheaton, Md., W1XG in Boston, and W8XF in Pontiac, Mich. In 1933 and 1934 several additional vhf stations were licensed to use the bands 42–56 and 60–86 Mc. In 1936, all activity in the 2-Mc band ceased and all vhf stations were placed in the bands 42–56 and 60–86 Mc.

In 1937, the frequency allocation was set up for the first time on the basis of channels 6 Mc wide. Nineteen such channels were set up between 44 and 294 Mc. The present allocation comprises 12 channels, each 6 Mc wide, in two groups 54–88 Mc and 174–216 Mc.

Evolution of High-Definition Standards

Shortly after permission to operate in the vhf bands was given, attention was focused on purely electronic methods of scanning and cathode-ray tubes for reproduction. Dr. Zworykin had demonstrated a cathode-ray receiver before The Institute of Radio Engineers in Rochester, N. Y., in November, 1929. In 1932, an "all-electronic" system was demonstrated by RCA, transmitting 240-line images from New York to Camden, near Philadelphia, over an air-line distance of about 80 miles, with one intermediate relay point at Arney's Mount, N. J. This was one of the earliest demonstrations of cathode-ray equipment, but the term "all-electronic" was something of a misnomer, since no satisfactory electronic synchronizing circuits had been developed, and the synchronizing pulses were derived by passing light to a photocell through apertures in a whirling disk. In 1934 P. T. Farnsworth announced a new electronic camera tube, the image dissector.

In 1934, the march toward higher definition got properly under way. Each transmitter was free to employ any scanning method, but between 1932 and 1934 agreement was reached that interlaced transmission, based on the "odd-line" principle, was the simplest and most satisfactory method of avoiding flicker in the reproduced images. The first "odd-line" value chosen was 343, an odd number composed of odd factors ($343 = 7 \times 7 \times 7$). From this root sprang many other choices, all tending toward greater definition in the pictures, all odd numbers, composed of odd factors. The majority opinion was that no more than 441 lines could be accommodated in a picture sent by double-sideband methods within the limits of a 6-Mc channel. The dissenting opinion was that it would be better to err on the high side in the number of lines, with consequent excessive definition in the vertical dimension, in the hope that better utilization of bandwidth would be possible as time went on. This dissenting opinion was in fact justified in 1939, when vestigial-sideband transmission was proved feasible and adopted as standard. The 441-line figure then proved too small for a 6-Mc channel and the standard was eventually changed (in 1941) to 525 lines, the value presently specified in the FCC regulations.

The foregoing paragraphs may indicate that the adoption of standards in the United States was accompanied by not a little dissension in the ranks. This must be admitted. In fact, the five-year period from 1936, when official sanction for public programs was given in the United Kingdom, to 1941, when similar sanction was granted in America, was characterized by a very vigorous debate on standards.

The debate finally came to an end in the meetings of the nine panels of the National Television System Committee (NTSC). This group of 168 television specialists, in the period from August, 1940, to March, 1941, devoted 4,000 man-hours to meetings, witnessed 25 demonstrations of the comparative merits of different proposals, and finally left behind them a record of reports and minutes some 600,000 words in length. Out of this monumental effort came virtually complete agreement on a set of 22 standards which were presented to the Federal Communications Commission for approval and adoption. This approval was granted, and commercial operation of television broadcast stations was authorized, to be effective July 1, 1941. The stage was set for a rapid advance. On December 7, 1941, the United States entered the war and the state of the art was frozen by lack of man-power and materials. Thereafter, until 1945, commercial broadcasting continued, but at a "bare-subsistence" level. The FCC required a minimum of 15 hours per week of public programs from each station before the war; after Pearl Harbor this was reduced to four hours per week.

In 1944–1945 a thorough review of the standards was conducted by the Television Panel of the Radio Technical Planning Board (RTPB). The standards were reaffirmed by this group, and no change has since occurred.

Evolution of Program Service

Prior to 1936, the American public was an incidental partner to the television enterprise. But in that year, the broadcasting of programs especially designed for public consumption began, although no regular source of receivers was available and no official sanction had been given for other than experimental transmissions. The occasion was one of considerable international competition between England and the United States. The first move came from the British Broadcasting Corporation. Plans for the opening of the London station in Alexandra Palace were announced early in 1936. This was the cue for the RCA-NBC transmitter to "get busy." On June 29, 1936, RCA started a field test of its television system with 100 of its engineers in and around New York as the observers. The images were transmitted by double sideband, at 343 lines, 30 frames per second. Film was used copiously in the early stages. But by November 6 of that year, the *New York Times* announced "the first complete program of entertainment over the NBC system," viewed enthusiastically by the press. Four days before, on November 2, the Alexandra Palace station had begun a regular public service. The American service was not public in the same sense, and did not become so for five years.

By 1938 the RCA field test had progressed to the point where its directors were ready to take a step in the direction of inviting the public to participate actively. It was announced in October, 1938, that, coincidentally with the opening of the New York World's Fair in May, 1939, regular public service would be offered.

In the meantime, vestigial-sideband transmission had been adopted as standard by the Television Committee of the Radio Manufacturers' Association and had been incorporated in the NBC transmitter. The scanning

pattern had been increased to 441 lines, and the effective video band to 4 Mc. In February, 1940, the Federal Communications Commission adopted rules permitting "limited" commercial operation of stations.

The stage seemed set. The NBC transmitter was offering 10 to 15 hours of program a week, including elaborate dramatic presentations from the studios, regular outside rebroadcasts of sporting events of every description, educational programs and films. But in April, 1940, the FCC retracted the offer of limited commercialization, following an announcement by RCA that receivers would be offered to the public in greater volume and at reduced prices. The FCC stated that this action by RCA tended to freeze the then accepted standards, those formulated by the Radio Manufacturers Association, without official sanction from the Government. This action effectively held up further progress until the standards had been studied and essentially reaffirmed by the National Television System Committee. In July, 1941, the impasse was cleared, and television broadcasting for the public officially began.

The American Television Channel

The standard 6-Mc channel assigned to television broadcast stations in the United States is intended for vestigial-sideband transmission. At the lower frequency limit of the channel, the emission is required to be substantially zero, actually no more than one per cent of the picture-carrier amplitude. At a point 0.5 Mc higher in frequency, the sideband emission has full amplitude. The picture carrier itself is located 1.25 Mc above the lower channel edge, and occupies an asymmetrical position with respect to the channel limits. Thus only a portion of the lower sideband is transmitted, hence the term "vestigial-sideband transmission." The upper sideband is transmitted fully over a region 4 Mc wide, i.e., it maintains maximum amplitude to a point 5.25 Mc above the lower channel edge. At this point the sideband energy is attenuated with increasing frequency until, at a point 5.75 Mc above the lower channel edge, it is attenuated to one per cent of the carrier amplitude. The carrier of the associated sound transmission is placed at this point. The remaining 0.25 Mc of the channel is reserved as a guard band.

This arrangement of carriers and sidebands was originally devised in 1938 and has persisted substantially without change through the deliberations of the National Television System Committee and those of the Radio Technical Planning Board. The vestigial-sideband principle permits a maximum unattenuated vision frequency of 4 Mc and permits attenuated transmission of vision signals up to a maximum of 4.5 Mc. In comparison, the double-sideband system permits a maximum vision frequency of 2.5 Mc. The vestigial-sideband transmission thus offers an increase in pictorial detail of 80 per cent, with substantial improvement in picture quality.

The wide spacing between carriers (4.5 Mc) was chosen in preference to the alternative narrow spacing (1.25 Mc) which would be possible if the sound carrier were transferred from the high frequency edge of the channel to the low frequency edge. The wide spacing produces a high-frequency beat frequency which is outside the limits of the vision frequency band and hence has little or no visible effect. The narrow spacing would produce a beat note within the video band.

The vestigial-sideband system is intended to operate with a receiver whose response characteristic increases linearly with frequency from 0.5 to 2.0 Mc above the lower edge of the channel. This "slope" region corresponds to the portion of the transmitter spectrum where both sidebands are transmitted. By virtue of the sloping receiver response the picture carrier is attenuated to one half, and the sum of the two sideband voltages is constant throughout the region, and equal to the value of sideband voltage at higher frequencies outside the "slope" region. Thus the video-frequency voltage developed at the output of the demodulator is the same for all sideband frequencies.

The ratio of sound-carrier power (radiated by the sound radiator) to picture-carrier power (radiated by the picture radiator) has been set up within the limits 0.5 and 1.50. The general range has been chosen to provide approximately equal areas of coverage of the sound and picture signals. The sound power is lower than the picture power because the sound transmission employs frequency modulation, with an inherent signal-to-noise ratio superior to that of amplitude modulation. Account has been taken, in setting up this ratio, of the fact that interference (particularly impulsive noise) in the sound channel is usually more objectionable than interference arising from the same noise source in the picture channel.

The Scanning Specifications

The standard American television picture is scanned in 525 lines from the beginning of one frame to the beginning of the next. Each frame is broken up into two fields of 262.5 lines each. The half-line portion at the end of a field causes the lines of one field to fall between the lines of the previous field. Hence an odd number of lines was chosen to give this half-line relationship between fields. The value 525 consists of odd integral factors ($525 = 7 \times 5 \times 5 \times 3$). This permits multivibrators or counting circuits in the synchronization generator (used to divide from the line frequency of 15,750 cps to the frame frequency of 30 cps) to operate in the most stable condition.

The frame frequency is 30 per second interlaced two-to-one. The aspect ratio of the scanning pattern (ratio of width to height) is 4/3, to agree with the ratio previously adopted as standard for motion-picture projection. The FCC standards specify that the active scanning of the picture shall occur at uniform velocity from left to right horizontally and top to bottom vertically.

The choice of 525 lines was made from among several proposed values, including 495 and 507 lines. Assuming 4.25 Mc as the maximum usable video frequency, equal vertical and horizontal resolution at 30 frames per second is obtained with a scanning pattern of about 500 lines, which would indicate that all the proposed values are equally suitable. The number 507 has the disadvantage of two large integral factors which require two of the frequency-dividing circuits in the synchronizing signal generator to count by a factor of 13. The choice between 495 and 525 was finally made on the basis of fineness of line structure, which indicates a slight preference for the higher number of lines. The value of 441 lines was discarded as it did not make full use of the maximum available video frequency.

Sound Signal Standards

A major difference between British and American television practice lies in the method of modulation employed for the sound transmission. The American standard specifies frequency modulation with a maximum frequency deviation, corresponding to the maximum audio level, of 25 kc either side of the unmodulated carrier frequency. Frequency modulation, employing a spectrum considerably wider than that required for the corresponding amplitude-modulated signal, has been shown to offer a substantially higher signal-to-noise ratio than that offered by amplitude modulation. This advantage obtains over all types of noise, provided only that the peak signal voltage is at least twice the peak noise voltage. Since natural atmospherics are rarely present in the vhf spectrum, the principal advantage of frequency-modulated transmission is the mitigation of impulse noises such as is generated by automobile ignition systems, and noise generated in tubes and circuit elements of the receiver.

To assist in the reproduction of the upper register of the audible spectrum, it has been customary in frequency-modulated sound transmissions to introduce audio pre-emphasis at the transmitter. The standard pre-emphasis characteristic is that of a series resistive-inductive impedance whose time-constant (resistance times inductance) is 75 microseconds. The converse de-emphasis is inserted in the receiver (usually by a resistive-capacitive impedance of the same time-constant). The advantage of such pre-emphasis lies in the fact that the sound power associated with the higher register is generally lower than that of the lower register and hence is less efficiently transmitted with respect to the noise level. Artificial emphasis and de-emphasis thus add to the over-all signal-noise ratio, by reducing the noise level in the upper register.

The remaining standard is the direction of polarization of the electric vector of the radiated wave. This was chosen as horizontal as early as 1938, and although polarization has been the subject of intensive investigation by the NTSC and the RTPB the advantage of the horizontal direction has been consistently upheld.

Rules Governing Allocation of Television Broadcasting Facilities

As the demand for broadcasting facilities has consistently exceeded the supply, it has been necessary for the FCC to set up equitable rules whereby the available portions of the spectrum may be allocated to serve the public interest.

A conflict in allocation arises, by definition, when interference occurs among stations. The interference is defined in terms of (1) the signal level required to give satisfactory service in an area from the station serving that area, and (2) the level of signal which creates interference in that area, arising from another station on the same channel in an adjacent area. The problem of interference between stations in the same area but assigned to adjacent channels must also be considered.

The basic level of service is defined as a field strength which must be equalled or exceeded over 50 per cent of the distance along a radial line from the transmitter. The field strength thus specified for built-up city areas and business districts is 5 millivolts per meter. For residential and rural areas the specified field strength is one-tenth as great, or 500 microvolts per meter. These figures properly surpass the figure of 50 microvolts per meter commonly regarded by engineers as the lower limit for "marginal service," in the absence of man-made sources of noise.

The applicant for a television construction permit or license must show that his proposed transmitter will offer service in accordance with the above rules, and must estimate the population lying within the 5-mv/m and 0.5-mv/m contours.

The interference ratio (ratio of desired signal to undesired signal) which must be equalled or exceeded within the service area has been set at 100:1 for stations on the same channel and 2:1 for stations on adjacent channels. This is in keeping with the commonly-held engineering opinion that an interfering signal must be at least 40 db below the desired signal if it is to have negligible effect. For adjacent-channel interference, the selectivity of the receiver circuits will introduce sufficient additional rejection if the interfering signal voltage does not exceed one-half the desired signal voltage. It is the practice of the FCC to avoid assigning adjacent channels in the same metropolitan area; otherwise the 2:1 ratio would be met in but a small portion of the normal service area.

The problem of finding sufficient facilities for television without interference is most critical in the highly populated areas along the eastern seaboard. In general an allocation plan to provide sufficient service to the cities of Boston, Providence, Hartford, New York, Philadelphia, Baltimore, and Washington will take care of all other population centers in North America.

Postwar Activity

The postwar period has been marked by a very substantial expansion of the television service. At present

(December, 1950) the areas covered by television signals include 63 major population centers, in which 107 stations offer service to a potential audience in excess of 50,000,000. The expansion has emphasized the shortage of spectrum space for television stations and has brought into sharp focus the problem of interference between stations.

In 1945, at the close of the war, about 10,000 receivers had been produced and sold to the public. Nine stations were operating in five cities, four of them on a fully commercial basis. Postwar production did not get under way, owing to shortages of materials and components, until the spring of 1946. In that year, 6,500 receivers were produced by the six manufacturers then in the business. Three years later (1949), the production was 2,400,000 sets, according to figures compiled by members of the Radio Manufacturers Association. Production by non-RMA companies accounted for approximately 450,000 additional receivers in that year. Over 100 companies participated in the 1949 production. Production in 1950 was approximately 6,000,000 receivers.

The trends in receiver design have been toward larger pictures, fewer tubes and simplification of controls. The smallest screen size is $2\frac{1}{2}$ inches in diameter, offered in a 1948 model no longer in production. The largest direct-view tube was an all-glass tube of 20 inches diameter, now replaced by a 19-inch metal-cone tube. From 1946 to 1948 the most popular model was the 10-inch set, but in 1949 this model gave way to the $12\frac{1}{2}$-inch size. The 16-inch tube took first place in 1950.

Projection pictures ranging from 9 by 12 inches to 18 by 24 inches have been offered in domestic receivers, although production of this type of receiver has been small compared with the direct-view type. Projection equipment for large audiences, ranging in screen size from 6 by 8 feet to the full theater screen of 18 by 24 feet has also been produced.

Reduction in the tube complement of receivers has averaged about 25 per cent since 1946. The first large-scale-production chassis was a 10-inch model employing 30 tubes. In 1948, a 7-inch model was offered with 16 tubes. Current models employ from 22 to 24 tubes. Substantially all receivers having screen diameter greater than 7 inches employ magnetic deflection of the picture tube.

Simplification of controls has resulted from improvement in the stability of circuits and components, particularly those used for tuning and for synchronization of scanning. The majority of receivers currently in production employ the intercarrier circuit, in which a sound intermediate frequency of 4.5 Mc is developed by frequency conversion at the picture second detector. The most popular value of picture intermediate frequency is 25.75 Mc, but plans for changing this to 45.75 Mc have been announced by many manufacturers and the latter value may eventually be adopted as standard by all manufacturers.

Expansion of Broadcast Facilities

By the end of 1946, 10 stations were in operation, and in that year 38 construction permits for additional stations were granted by the FCC. By the fall of 1948, 54 stations were operating, construction of an additional 70 stations had been authorized, and 310 applications for construction permits had been filed.

Beginning in the spring of 1948, the FCC received complaints that excessive interference was occurring between stations on the same channel, as well as between stations occupying adjacent channels. The co-channel interference takes the form of horizontal bars, which move upward or downward through the image of the desired station. These bars, known as "venetian-blind" interference, are caused by the heterodyne beat between the carriers of the stations. The beat frequency can have values from a few cycles per second to several thousand cycles per second, under the frequency tolerances permitted by the FCC regulations.

The interference extends over an appreciable portion of the service area of each station, for a majority of the operating hours, when the stations were separated by less than approximately 200 miles. It is explained by tropospheric propagation which produces signal strengths much higher than had been anticipated when the allocation plan was devised. Since the then-existing allocation plan called for a minimum spacing of 150 miles, and exceptions to the rule as small as 95 miles had been allowed, the FCC decided to call a halt on further expansion of broadcasting facilities until the cause and extent of the interference could be studied. Accordingly, in September, 1948, the Commission issued a "freeze" order, prohibiting construction of new stations for an indefinite period. The order did not prohibit completion of the 70 stations then under construction, but held up any further action on the 310 applications for construction permits. The freeze order was still on the books, as of December, 1950. At present nearly 400 applications for new stations are on the books of the FCC.

Meanwhile, the means of accommodating this large number of stations, while minimizing interference, have been under almost continuous study in a series of conferences and hearings before the FCC, from September, 1948, to the present. Two approaches have been investigated: (1) the reduction of co-channel interference by control of the interfering carriers, and (2) the extension of the television spectrum into the ultra-high frequencies.

Two techniques of interference reduction have been developed, carrier synchronization and carrier-offset operation. The second of these is now in wide use among co-channel stations in the east and midwest sections of the country.

In the carrier synchronization scheme, the two carriers are received in a monitor station approximately midway between the stations, and their frequencies compared

in a frequency discriminator. A signal representative of the frequency difference is sent over a telephone line to one of the stations, where it actuates a control circuit connected to the quartz-crystal frequency control. By this means, the carriers of the two stations are kept rigidly locked together in precise phase relationship. Since the beat frequency between the carriers is thereby reduced to zero, the venetian-blind interference is removed. The strength of the interfering signal may then be allowed to increase by about 17 db on the average before the image content of the undesired signal becomes objectionably apparent. This reduction in interference would permit co-channel stations to be located at a distance of separation of about 150 miles, compared to about 210 miles for the same degree of interference with carriers unsynchronized.

The cost of the monitoring station and the telephone lines necessary in the carrier synchronization method led to the development of a simpler system known as carrier offset. In this method, the carrier of one co-channel station is purposely separated from that of the others by 10.5 Kc, plus or minus the 0.002 per cent tolerance allowed by the FCC regulations. The average beat frequency of 10.5 Kc produces a large number of venetian-blind bars, each occupying about three adjacent scanning lines. Since each interference bar is thereby confined to the space occupied by two successive lines in a single field scanning, the average tone produced tends toward a neutral gray, and the effect is subjectively much less annoying than if the carriers were within a few hundred cycles of one another, as is usually the case between unsynchronized carriers.

Actually, the greatest reduction of interference is obtained when the offset frequency is 7.875 kc, that is, one-half the line scanning frequency, and no improvement is noted when the offset is 15.750 kc, the line scanning frequency. The half-line frequency would be used in the offset system, were it not for the fact that three or more co-channel stations are often involved in a single interference area. The insertion of 7.875 kc offset between all stations would then result in nearly zero or 15.750 kc offset (no interference reduction) between certain pairs in the group. The 10.5-kc figure was chosen as a compromise, permitting several stations to operate with carrier offset in the same area. The average interference reduction, relative to the unsynchronized case, is about the same as the carrier synchronization method. No wire line or monitor is needed; the crystal control frequency of a particular station is merely chosen to produce a carrier 10.5 kc higher or lower than that of the neighboring stations.

The investigation of ultra-high frequencies for television broadcasting has centered on the frequency range from 475 to 890 Mc, a region of the spectrum previously set aside by the FCC for experimental television stations. In planning to turn over all or part of this band to commercial broadcasting, several decisions have to be weighed: (1) what standards of transmission to adopt, including the question of whether the service should be in black-and-white, in color, or both; (2) the bandwidth of the channels, and consequently the number of channels available in a given portion of the band; and (3) whether to assign ultra-high-frequency (uhf) stations to cities already possessing vhf stations, or to separate the two services geographically as much as possible.

Since the demand for additional black-and-white stations is so great, it has been conceded that at least part of the uhf band must be turned over to the black-and-white service. Moreover, to make the additional service available to owners of existing receivers at the least expense, the standards of transmission, including the channel width of 6 Mc, should be identical to those of the vhf service. This latter requirement exists even if the uhf stations are to be confined to cities not now served by vhf stations, since owners of existing receivers may move from a vhf city to a uhf city.

This leaves unsettled the question of whether color service should also be provided in the uhf band. Prior to 1948, all planning for commercial color television was on the basis of channels 12 to 16 Mc wide, and there seemed to be no room for a national allocation of such channels, even if the whole of the uhf band were used. But since that time color service on 6-Mc channels has been authorized, as outlined below, and it is now planned that color transmissions be offered by black-and-white stations on 6-Mc vhf and uhf channels.

The third question, whether or not to mix uhf and vhf assignments in a given locality, arises from the fact that the propagation conditions affecting uhf service indicate that the primary service area of a uhf station would be appreciably smaller than that of a vhf station of equal radiated power. Field tests conducted in Washington, D. C., and Bridgeport, Conn., have confirmed this assumption, at least to the extent of indicating that the shadowing effect of obstructions on uhf channels is much sharper than on vhf channels. Out to perhaps 25 miles, the service on the two bands is expected to be nearly equivalent, except for shadow effects, but beyond this limit, and out to the radio horizon, the vhf station would provide a better grade of service.

If these preliminary findings hold true generally, the licensee of a uhf station would find himself at a competitive disadvantage with respect to the licensees of vhf stations in his locality, and this would be particularly true if the potential audience of the stations extends in any substantial density beyond the 25-mile limit. In such cases, it may be questioned whether broadcasters would wish to make the investment in a uhf station.

On the other hand, if no vhf assignments are made in certain cities, the potential viewers located beyond the range of a uhf station might be denied service which otherwise could be rendered by a vhf station.

The uhf allocation proposed by the FCC in the Fall of 1949 envisages 42 additional 6-Mc channels, starting at 475 Mc or 500 Mc, and extending upward to 727

Mc or 752 Mc. Vhf and uhf stations are assigned together in certain localities in this plan. The FCC proposal was made simply as a basis for discussion in the hearing which is currently under way. Detailed argument on the proposal began in October, 1950.

Theater Television

Throughout the development of television as a medium of home entertainment there has existed a parallel effort directed toward a system suitable for large audiences in theaters. The earliest demonstration of large-screen television was conducted by E. F. W. Alexanderson in Schenectady in 1929, using the scanning disk method to project 60-line images. But the low optical efficiency of this system did not permit a sufficiently bright picture to be projected when the number of lines was increased to 525.

Attention then turned on the method of projecting the image formed on the face of a cathode-ray tube. Since large lens apertures were required, refractive projection lenses proved exhorbitantly expensive for projecting images of theater size. A more efficient system using reflective optics, based on the Schmidt telescope, was finally adopted as the basis for direct projection. In the Schmidt system, the image is reflected from a spherical mirror through an aspherical correction plate which removes spherical aberration. The first demonstration of a Schmidt projector was held by RCA at the New Yorker Theater in New York on January 24, 1941. The equipment employed a seven-inch picture tube with a second anode voltage of 70,000. A highlight brightness of about 5 foot-lamberts was achieved on a 15-by-20 foot screen.

In an attempt to achieve greater brightness on larger screens (film projectors produce about 10 foot-lamberts on screens as large as 18 by 24 feet), the intermediate film method was developed. In this process, the image on a picture tube is photographed on motion picture film, which is processed immediately and passed through the conventional film projector. The processing time has been reduced to less than one minute, so the loss of immediacy is not important, except possibly in sporting events. The film serves as a permanent record of the performance and permits repeat showings of programs. The first demonstration of the intermediate film method was given at the Paramount Theater in New York on April 14, 1948. This equipment has since been used to televise to the theater audience such events as the national political conventions in 1948 and the Joe Louis-Joe Walcott prize fight.

Color Television

The recent history of color television in the United States starts in 1940, when the Columbia Broadcasting System, demonstrated a 343-line 120-field system employing sequentially-scanned fields in three primary colors. The system as then proposed used a 6-Mc channel, and experimental broadcasts were made in the 50-to-56-Mc channel then occupied by the CBS black-and-white transmitter. This work was interrupted by the war. Later, in 1946, the standards were changed to 525 lines, 144 fields, to provide higher resolution and greater freedom from flicker. These values required a channel 16 Mc wide. To bring the channel width more in line with existing 6-Mc black-and-white service, the channel was later reduced to 12 Mc and the number of lines was reduced to 441. The 144-per second field rate was retained to avoid flicker. This system was proposed by CBS for commercial use in the ultra-high frequencies in 1946, but in March, 1947, the FCC denied the CBS petition on the ground that art had not advanced sufficiently to justify adopting standards at that time.

By 1948, the CBS system had been shifted to a 6-Mc channel, employing 405 lines, 144-fields, but otherwise similar to the previous proposals. At first this system was shown only on a "closed circuit" basis (i.e., not broadcast) but in the summer of 1949 the FCC asked for information concerning the possibility of immediate introduction of color service on 6-Mc channels, and the CBS proposed their system for this purpose.

The CBS system is known as the field sequential system, since the color sequence is introduced by changing the color at the completion of each scanning field. The high field rate of 144 per second is necessary to avoid flicker at image brightness (in the highlights) up to about 25 foot-lamberts. Since the system uses the same bandwidth as the black-and-white system, the total number of resolvable picture elements in the color image is in the inverse ratio of the field scanning rates, or $60/144 = 0.42$, that is, 42 per cent of the resolution of the standard black-and-white resolution image. This loss of resolution, plus the fact that existing receivers would have to be converted to new line- and field-scanning rates to permit reception of the color signals, have been urged as disadvantages of the field-sequential system.

As early as 1946, the desirability of providing color service on scanning standards identical to those of black-and-white service, thus minimizing obsolescence of existing equipment, was manifest. In that year the Radio Corporation of America announced the so-called simultaneous system, in which three entirely separate and complete images, each scanned at 525 lines, 60 fields, were produced in the three primary colors by three camera tubes, in optical and electrical register. For color service, the three signals, radiated in three adjacent subchannels occupying a total band about 15 Mc wide, were applied to separate picture tubes, and the images combined, in register, by projection on a common viewing screen. The black-and-white version of the color transmission would be available to existing receivers by tuning to the green channel which contains most of the tonal information of the color image. Since this channel, as well as the red and blue ones, is based on 525-line, 60-field scanning, no change in the equipment, beyond a uhf antenna and radio-frequency converter to

tune to new channels, would be required. Later a technique known as mixed highs (which allowed reduction in the bandwidth required for the red and blue images) permitted this system to operate on a 12-Mc channel. When attention was directed, in 1948, toward the 6-Mc color service, the simultaneous system was discontinued, and work on the dot-sequential system was instituted.

The dot-sequential system introduces the color sequence by changing the color between successive picture elements (dots) along each line. This permits a second interlace between dots to be added to the conventional line interlace. A color image transmitted by this method at the black-and-white standard of 60 fields per second has a flicker performance equal to that of the black-and-white system, that is, high light brightness above 100 foot-lamberts is permissible without evident flicker. The image structure is divided among the sets of interlaced lines and the sets of interlaced dots in such a way that all points on the image are scanned in all colors 15 times per second.

The dot sequential system is particularly adapted to transmission by the time-multiplex method, which is considerably more efficient in its use of the spectrum than is the continuous modulation method employed in black-and-white system and the other proposed color systems. In fact, time-multiplex transmission provides nearly two-to-one improvement in resolution for a given bandwidth, i.e., the image detail is equal to that provided by continuous modulation over an 8-Mc band, when transmission occurs by time multiplex over a 4-Mc band. By this technique it is possible to transmit a color image, within a 6-Mc channel, having nearly as high resolution, and equal freedom from flicker, as a black-and-white image transmitted by continuous modulation over the same channel. Moreover, the scanning rates of the dot-sequential color system are identical to those of the black-and-white system so obsolescence of existing equipment is avoided.

The time multiplex transmission is achieved by sampling in sequence the outputs of three camera tubes (which televise the scene simultaneously in the three primary colors) at a rate of 3.6 Mc. Sinewaves in three-phase relationship are developed from this sampling process, each representative of the scanning of alternate dots along the lines of the respective images. The sinewaves are combined vectorially into a single sinewave which is radiated as a sideband component separated 3.6 Mc from the picture carrier. At the receiver, the combined sinewave is sampled by an electronic switch synchronized with the sampling switch at the transmitter (the synchronization is accomplished by short bursts of 3.6 Mc sinewave inserted on the sync pedestal during the horizontal blanking period).

The separated sinewaves resulting from this process control three electron guns in a tricolor picture tube. The viewing screen of the tricolor tube consists of several hundred thousand dots of phosphor material arranged in groups of three. One phosphor dot in each group fluoresces with red light, the second with blue light, and the third with green. The area of the three phosphor dots as a group corresponds to the area of a single picture element. Accordingly, each picture element may be caused to assume the desired combination of primary colors by sequential excitation of the phosphor dots within its area. The three electron guns are so positioned that the electron beam of one gun can excite only the red phosphor dots, the second gun only blue dots, and the third only green dots. A single-gun type of tube, in which the beam traverses a spiral in impinging on the screen and hence can excite one phosphor dot to the exclusion of the other two in the group, has also been demonstrated.

The terminal apparatus (camera and picture-reproducing equipment) of the dot-sequential system is more complicated than that of the field-sequential system. Moreover, it is vital that the transmission system cover the full video frequency range up to and including the sampling frequency of 3.6 Mc; if this frequency is not transmitted, the color values are lost and a rendition in tones of gray results.

The third color television system considered by the FCC is the line-sequential system developed by Color Television, Inc. In this system the color sequence is introduced by a change in color at the end of each scanning line. The scanning rates are 525-lines, 60-fields, as in the black-and-white system and dot-sequential color system. The color sequence is obtained by focusing on the mosaic of an image orthicon, three congruent images side by side, in the three primary colors. The group of three images is scanned at one third the usual line scanning rate ($15,750/3 = 5,250$ cps). Each horizontal passage of the scanning beam crosses three images, so the lines are scanned at 15,750 cps in the sequence red, green, blue. Since the number of lines per frame (525) is exactly divisible by three, it would follow that a given line in the scanning pattern would always be scanned in the same color, and the rendition of solid tones would display very poor vertical resolution. To avoid this effect, the colors are commutated by a special sync signal which causes a vertical shift of certain lines. In this manner every line is scanned in three colors at a rate of 10 per second. This rate is sufficiently low to create interline flicker and line crawl.

On October 11, 1950, the FCC announced its decision to adopt the CBS field-sequential color system, and stated that it would authorize commercial color broadcasts using this system on and after November 20, 1950.

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Analysis of Synchronizing Systems for Dot-Interlaced Color Television*

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Summary—A mathematical analysis is made of two synchronizing systems which might be used in "dot" color television to synchronize the dotting or sampling frequency in the receiver. Synchronizing information is transmitted in bursts of carrier cohered in phase of approximately 3 Mc during line fly-back time. The two systems analyzed are (1) a simple high-Q resonant filter and (2) an oscillator with automatic frequency control (afc).

In order to maintain sufficient phase accuracy in the sampling frequency, crystal control at the transmitter is necessary. Since the variations of the frequency response of a particular crystal filter with time may be made essentially negligible, the problem resolves itself into calculating the parameters of the synchronizing system to keep below an acceptable value the random phase error due to noise together with the phase error caused by variations in frequency (static phase error) from crystal to crystal. Rather than attempt to minimize the sum of the random and static phase errors, the two have been dealt with separately because of their different character. A fundamental parameter in the calculations is the power carrier-to-noise ratio in the intermediate frequency, the carrier being measured at the sync tips. Calculations are made for values of this ratio of 1, 3, 5, this being considered the critical range.

Results of the calculation show that when the phase error due to noise alone is fixed, the simple resonant filter and the afc with single time constant suffer the same static phase error. If phase errors in the neighborhood of 10° can be tolerated, it appears then that receivers can be designed to operate without manual control of the dotting frequency down to carrier-to-noise ratios of about 2. This should be satisfactory since picture quality at this level is very poor.

IN DESIGNING a "dot interlace" system[1] of color television, synchronizing problems over and above those of ordinary black-and-white television are encountered. A "dot" system consists essentially of a three-channel time division multiplex system with one channel devoted to the transmission of each primary color. This involves sampling each color at the transmitting end of the system at approximately a 3-Mc rate, transmitting information so derived to the receiver, separating this information by resampling at the same 3-Mc rate and applying it to control the three colors displayed on the cathode-ray tube. Thus a sample of each color is obtained on every third pulse. In order to avoid deleterious color distortion, the sampling frequency in the receiver must be closely synchronized in phase with that of the transmitter.

In the receiver we have the problem of first synchronizing the frame, then the line, and finally the dots within each line. The first two of these are common to both color and black-and-white television and since these problems are relatively well understood, less attention will be given to them than to dot synchronizing. There appear to be no smaller tolerances required for frame and line synchronizing in color television than in black-and-white. A continuous-wave 3-Mc wave could be transmitted, picked out by frequency separation, and used for synchronizing purposes. This has obvious disadvantages, however. A better scheme, the one which will be considered here, consists in transmitting a burst of 3-Mc carrier during horizontal sync pulse. This burst together with the sync pulse may be separated from the video in the usual way by clipping. The 3-Mc information may then be separated from the pulse by filtering or by gating. Alternatively, one could first gate out the 3-Mc burst and then filter. The clipped pulse and burst may be applied directly to the conventional pulse afc system without difficulty. However, the pulse must be removed in some way before the burst is applied to a sine-wave afc system.

Assuming that the 3-Mc burst has been in some way separated, it is necessary to consider what mechanism will be used to effect synchronization. Since the tolerances on the stability of the transmitted synchronizing wave will largely determine the mechanism to be used, these will be established. It will be assumed that all synchronizing frequencies in the transmitter are crystal-controlled so that it may be expected that variations in the 3-Mc frequency will not exceed ± 50 cps when switching from station to station. Furthermore, the drift in frequency of a given crystal will be so small as to be negligible. Also, the ratio between the dotting frequency and the line frequency is so chosen that there is no necessity to shift the phase of the dotting frequency at the end of each line or frame. Under these conditions three possible synchronizing devices appear possible: first, a simple high-Q filter (e.g., a crystal filter) which picks out the 3-Mc component, amplifies it, and applies it directly as the sampling wave; second, a locked sine-wave oscillator controlled in frequency by the 3-Mc burst; third, an afc system which controls the frequency of a sine-wave oscillator. Since any sync system must operate satisfactorily under bad fluctuation noise conditions, it must be realized that this requirement will ultimately determine the choice of system.

If, in order to achieve good noise protection, the simple filter device is narrowed down, the point will eventually be reached where the slope of the phase characteristic becomes so large that shifts in frequency such as may be encountered in switching from station to station will cause such an intolerably large static phase error that a manual frequency trimmer must be used. Furthermore, lock-in time on sudden phase transi-

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[1] W. P. Boothroyd, "Dot systems of color television," *Electronics*, vol. 22, pp. 88–93; December, 1949.

tions may be too long. This dilemma is common to any sync system in various degrees; that is, if the system is designed to ignore noise fluctuations it tends also to ignore variations from any source.

All three of the possible systems will track variations in phase and frequency with varying degrees of fidelity. The locked oscillator may have some advantage over the simple filter in that its effective bandwidth varies directly as the voltage of the applied synchronizing signal, although its chief merit appears to be in the fact that it is a Q multiplier.[2] Automatic-frequency-control sync would appear to be the most efficient of the three since it separates the functions of phase detection, filtering and amplifying, and generation of the correct synchronizing signal, thus permitting the parameters in each part of the system to be adjusted for optimum performance. The validity of these statements requires some proof and this will now be undertaken. The simple tuned circuit will be compared quantitatively with afc sync, since it is believed that locked oscillator performance lies somewhere between these two.

Since crystal frequency control is assumed used in the transmitter, the problem essentially reduces to the calculation of phase error introduced by random noise, and the static phase error caused by a shift of frequency when station switching. The delay in tracking a phase shift caused by station switching is not believed to be important.

The characteristics of each system which will keep the phase error due to random noise alone below 5° will now be determined, and then the respective static phase errors caused by transients of frequency will be found. This will be done for various signal-to-noise ratios in the intermediate frequency and will furnish a measure of the fidelity of the two systems. The average carrier power-to-noise ratio in the intermediate frequency will be used where the carrier is measured at the sync tips.

If the carrier burst occupies 4 μseconds out of 64 μseconds, when gated, clipped, and filtered, it will yield a sinusoid of amplitude $(C/128)\beta$ where C is the amplitude of the intermediate-frequency carrier during the sync pulse and β is the modulation suppression term derived in Appendix A.

The noise power out of a filter of total bandwidth 2γ will be $2\gamma F_0$ where F_0 is the amplitude of the noise-power spectrum at 3 Mc as derived in Appendix A and 2γ is the equivalent noise bandwidth of the filter. 2γ will turn out to be so small compared with the 4-Mc video bandwidth that the power spectrum may be considered flat across it. When noise is added to a sinusoid, the phase of the sinusoid becomes indeterminate to a degree. It is shown in Appendix C that in such a case, the rms value of the phase error due to noise alone is $\sqrt{\psi_0}/Q$ where ψ_0 is the total noise power added to the sinusoid, and Q is the amplitude of the sinusoid. Thus if it is desired that the rms value of the phase error due to noise alone be less than 5°,

$$\frac{\sqrt{2\gamma F_0}}{Q} < \frac{5}{57.3}.$$

Using the values of F_0 from Appendix A and considering the filter to be a single-tuned circuit it is found that the half bandwidth α must be less than 22, 85, and 135 cps, respectively, for intermediate-frequency signal-to-noise ratios of 1, 3, and 5.

To calculate the static phase error, the filter is again taken to be a single-tuned circuit and the maximum frequency deviation to be ± 50 cps. Then the slope of the phase characteristic at resonance is

$$\left.\frac{d\theta}{d\omega}\right|_{\omega=\omega_0} = \frac{1}{\alpha}$$

where α is the half bandwidth. Thus measuring frequency deviation from resonance, one has $d\theta = d\omega/\alpha$. This yields a phase shift of 2.3, 0.6, and 0.37 radians for intermediate-frequency signal-to-noise ratios of 1, 3, and 5. This amount of phase shift obviously requires an auxiliary manual phase control. If the allowable rms phase error is 10°, then the bandwidths become 88, 340, and 540 cps, respectively, and the static phase shift becomes 33°, 8.4°, and 5.3°. The picture probably becomes unusable for a carrier-to-noise ratio in the intermediate frequency of about 2, so that if these phase errors can be tolerated in the sampling frequency, then the receiver can be operated without auxiliary manual hold control. The transient caused by a sudden step of phase in the input to the filter will have the time constant of the filter. This will occur when switching stations, but it is not considered to be important.

To determine the characteristics of the afc synchronizing system, the configuration of Fig. 1 is assumed.

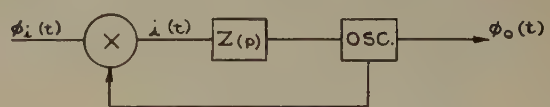

Fig. 1—Automatic-frequency-control servo loop.

The incoming sync information after detection and separation is fed into a phase comparator where its phase is compared with that of the free-running oscillator which determines the frequency of the receiver sync. Thus the incoming phase information is denoted by $\phi_i(t)$. Out of the phase comparator comes an error current $i(t)$ which passes through a filter $Z(p)$ before correcting the oscillator whose output is $\phi_0(t)$. Then out of the phase detector there is obtained

$$i(t) = K_1[\phi_i(t) - \phi_0(t)]$$

where K_1 is the sensitivity parameter of the phase comparator and is taken to be a constant. From the filter $Z(p)$ there is a voltage $e(t)$ which defined by

[2] Kurt Schlesinger, "Locked oscillator for television synchronization," *Electronics*, vol. 22, pp. 112–118; January, 1949.

$$e(p) = Z(p)i(p).$$

It is assumed that the voltage controlling the oscillator is held at a bias which yields the nominal sync frequency so that the voltage applied here results in deviation about that frequency. It is further assumed that the instantaneous deviation of oscillator frequency from the nominal value is directly proportional to $e(t)$.

Then

$$\frac{d\phi_0}{dt} = K_2 e$$

where K_2 is the sensitivity constant of the oscillator.

Since the phase comparator is a nonlinear device, K_1 is not an absolute constant but depends somewhat on the input signal and noise level. However, when measured or otherwise determined for a particular signal and noise, it will not vary much for reasonably small deviations from that value of signal and noise. Presumably, in designing an afc sync system one must first decide the lowest signal-to-noise ratio under which the system is to operate satisfactorily. The design is then optimized for that condition with the expectation that performance will be better when the noise level falls. K_2 will be essentially constant for small values of $e(t)$ for sine-wave oscillators controlled by reactance tubes or for blocking oscillators. Then, solving these equations in the frequency domain, it is seen that

$$\phi_0(p) = \frac{K_1K_2Z(p)}{p + K_1K_2Z(p)}\phi_i(p).$$

If $Z(p)$ is taken as a simple RC low-pass filter,

$$Z(p) = \frac{A_0}{p + \alpha}.$$

Then

$$\phi_0(p) = \frac{A_1}{p^2 + \alpha p + A_1}\phi_i(p)$$

where $A_1 = K_1K_2A_0$. Thus the over-all gain function of the system is given by

$$Z_0(p) = \frac{A_1}{p^2 + \alpha p + A_1}.$$

Now consider the effect on the system of introducing signals such as a step of phase or a step of frequency. These situations will exist when stations are switched. The output phase is given by a Laplace transform, thus

$$\phi_0(t) = \frac{A_1}{2\pi j}\int_{-j\infty}^{j\infty}\frac{g(p)e^{pt}dp}{p^2 + \alpha p + A_1}$$

where $g(p)$ is the spectrum of the input $\phi_i(t)$. In the case of a step of phase of magnitude I_0,

$$\phi_0(t) = \frac{A_1 I_0}{2\pi j}\int_{-j\infty}^{j\infty}\frac{e^{pt}dp}{p[p^2 + \alpha p + A_1]},$$

where

$$p^2 + \alpha p + A_1 = (p - p_1)(p - p_2)$$

and

$$p_1 = \frac{-\alpha + \sqrt{\alpha^2 - 4A_1}}{2}, \qquad p_2 = \frac{-\alpha - \sqrt{\alpha^2 - 4A_1}}{2}.$$

These roots are always in the left half plane and may be complex or real depending on $\alpha^2 - 4A_1$ being less than or greater than zero. In the former case, there is no oscillatory approach to the final tracking position while the latter provides a hunting approach. Operation at or near the critically damped position when $4A_1 = \alpha^2$ is generally preferred. In case the system is under-damped, the output is

$$\phi_0(t) = I_0\left[1 - \frac{e^{-(\alpha/2)t}\left(\frac{\beta}{2}\cos\frac{\beta}{2}t + \frac{\alpha}{2}\sin\frac{\beta}{2}t\right)}{\frac{\beta}{2}}\right],$$

where $\beta = \sqrt{4A_1 - \alpha^2}$. This will reduce to the critically damped case when $\beta = 0$. $\phi_0(t)$ can be written as

$$\phi_0(t) = I_0\left[1 - \frac{2\sqrt{A_1}}{\beta}e^{-(\alpha/2)t}\cos\left(\frac{\beta}{2}t - \tan^{-1}\frac{\alpha}{\beta}\right)\right].$$

The output then is as shown in Fig. 2.

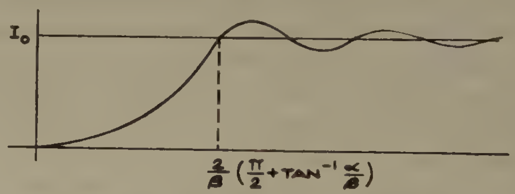

Fig. 2—Output of afc system to a step of phase.

The actual crossing of the line occurs at

$$t = \frac{2}{\beta}\left(\frac{\pi}{2} + \tan^{-1}\frac{\alpha}{\beta}\right)$$

so that if this crossing is taken as a measure of lock-in time, it appears that the lock-in time can be made small by making α large and β large. However, the overshoot becomes larger in such a case, so that a reasonable compromise must be made between fast lock-in time and initial overshoot. By minimizing the error $i(t)/\kappa_1$ in the mean square sense, it can easily be shown to vary as $1/\alpha$ so that in this sense, optimum tracking is obtained by making α as large as possible. All of this is in the absence of noise. It may be further noted that the output actually tracks I_0 regardless of loop gain or filter bandwidth and that tracking time is independent of the magnitude of the phase error. To maintain an oscillatory approach, $4A_1 > \alpha^2$. If for example $4A_1$ is taken to be equal to $2\alpha^2$ and $4\alpha^2$, respectively, the first overshoot is 4 per cent and 16 per cent, which gives an idea of the

magnitude of the oscillation. The magnitude of the first overshoot is $I_0(1+e^{-\pi\alpha/\beta})$. These oscillations have been observed experimentally.

Now suppose that the input signal is in correct phase but off in frequency by a fixed amount, due to oscillator drift or station switching. Then beginning at $t=0$, $\phi_i(t)=\omega_0 t$ where ω_0 is the frequency difference. Then $g(p)=\omega_0/p^2$ and

$$\phi_0(t) = \frac{2\omega_0}{\beta} e^{-(\alpha/2)t} \cos\left[\frac{\beta t}{2} - \tan^{-1}\left(\frac{\alpha^2 - 2A_1}{\alpha\beta}\right)\right] + \omega_0 t - \frac{\omega_0 \alpha}{A_1}.$$

This last function is asymptotic to $\omega_0 t-(\omega_0\alpha/A_1)$ so that there is a static phase error $\omega_0\alpha/A_1$. This static phase error results in a static shift in the voltage $e(t)$ applied to the oscillator which is just enough to change its frequency by ω_0. One notes that this error is ω_0/RK_1K_2 and is thus independent of the capacity of the filter. The output is as shown in Fig. 3.

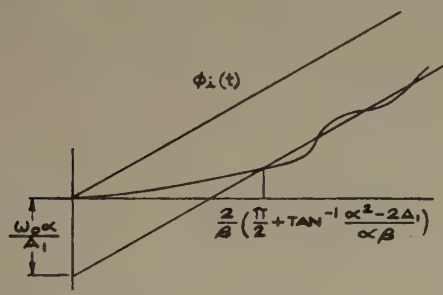

Fig. 3—Output of afc system to a step of frequency.

In general, in designing an afc system, it is necessary to consider the necessity of tracking relatively slow variations in the incoming signal in the presence of noise as well as the more severe transients such as steps of frequency and phase. Having decided upon a typical input signal to be tracked with optimum fidelity in the presence of a given noise level, an attempt would then be made to find the parameters which would minimize the error in some sense. This objective leads naturally to the theory of Wiener[3] in which an attempt is made to find the filter which will minimize the difference between input and output in the mean square sense. In principle, this procedure is straightforward. However, it will generally yield a filter characteristic which is not physically realizable and must therefore be approximated. It is therefore the practice of servomechanists to start with a servo of a given type having adjustable parameters. The mean square error can then be computed directly and the parameters adjusted to minimize it. For our purposes here the simple low-pass filter heretofore assumed is believed adequate. An additional degree of freedom with some improvement in performance can be obtained by starting off with a double time-constant filter. However, this complicates the calculations considerably and will not be gone into here.

The output of the servo as a convolution integral can be given as

$$\phi_0(t) = \int_0^\infty [\phi_i(t-\tau) + \phi_N(t-\tau)]K(\tau)d\tau$$

where ϕ_N is the noise assumed to enter the system with $\phi_i(t)$ and $K(\tau)$ is the characteristic transient of the servo system. The phase error is given by

$$\frac{i(t)}{K_1} = \phi_i(t) - \phi_0(t)$$

$$= \phi_i(t) - \int_0^\infty [\phi_i(t-\tau) + \phi_N(t-\tau)K(\tau)d\tau.$$

Assuming no error in the incoming signal, the phase error becomes

$$\frac{i(t)}{K_1} = -\int_0^\infty \phi_N(t-\tau)K(\tau)d\tau.$$

Thus using Plancherel's theorem, the mean-square phase error due to noise alone is

$$\frac{1}{4\pi}\int_{-\infty}^{\infty} F_1(\omega)\,|Z_0(\omega)|^2 d\omega$$

where $F_1(\omega)$ is the phase noise power spectrum and $Z_0(\omega)$ the over-all servo gain function. The above considerations are quite general and apply to any afc system of the type considered here. It may be assumed that the bandwidth of $Z_0(\omega)$ will be so small compared to 4 Mc that $F_1(\omega)$ can be considered flat across it. Therefore, the mean-squared phase error is

$$\frac{1}{4\pi} F_1 \int_{-\infty}^{\infty} |Z_0(\omega)|^2 d\omega = \frac{F_1}{4\pi}\int_{-\infty}^{\infty} \frac{A_1{}^2\alpha\omega}{(A_1-\omega^2)+\alpha^2\omega^2}$$

$$= \frac{F_1 A_1}{4\alpha}.$$

It may be observed here that in trying to minimize this error, the static phase error $\omega_0\alpha/A_1$ is made large so that a compromise must be struck between the two. Since this result will be evaluated for $4A_1=\alpha^2$, little error is caused here by taking the integral to ∞.

Now consider that the afc is applied to the 3-Mc carrier. First the burst is put through a narrow tuned circuit to pick out the 3-Mc sine wave. No phase information is lost by this maneuver. This filter must be wide enough so that an objectionable phase shift does not develop here. Then a sine wave plus noise is put into the phase detector. The power spectrum of the phase noise is, as shown in Appendix A,

$$F_1(\omega) = \frac{2F_0}{Q^2}, \qquad 0 < \omega < \gamma$$

[3] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," Radiation Laboratory Series, vol. 25, McGraw-Hill Book Co., New York, N. Y., chap. 7; 1947.

where Q is the amplitude of the sinusoid, 2γ the bandwidth of the prephase-comparator filter, and F_0 the power spectrum of the video noise at 3 Mc.

Now we wish to compare the simple filter with the afc by constraining them to have the same rms phase noise error and then comparing the ensuing static phase error. If the phase noise errors are equated, it is found that $\alpha_1 = A_1/\alpha_2$ where α_1 is the half bandwidth of the simple filter, and α_2 the half bandwidth of the filter in the afc loop. Then the static phase error for the afc case is $\omega_0\alpha_2/A_1 = \omega_0/\alpha_1$ which is identical with the result for the simple filter, independent of the relationship between A_1 and α_2. If a double time-constant filter is used in the afc loop, it can be shown that for equal static phase errors, the noise error in the double time-constant case can always be made less than that for the single time-constant filter. In such a case, there will then be improvement in performance over the case analyzed here at some reduction in lock-in time.

It will now be shown how the foregoing analysis may be applied to pulsed sync systems, although specific calculations will not be carried through, since such systems are not the primary interest here.

Consider the conventional black-and-white afc pulse sync system. Here the sync information is not provided continuously but intermittently so that there is a little difficulty in defining phase in a sense that may be compared with the continuous case. The position of incoming pulses is determined in a conventional balanced

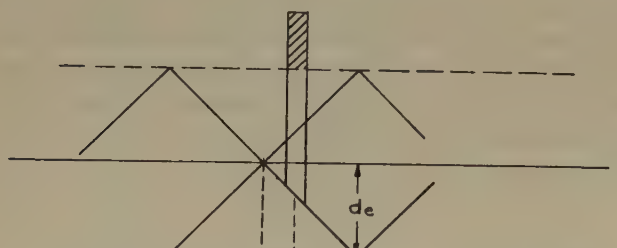

Fig. 4—Balanced pulse phase comparator.

phase detector[4] by feeding back part of the sawtooth output of the oscillator in the fashion shown in Fig. 4. In effect, the pulse is caused to ride on each of two sawteeth which differ only in polarity. In each case the pulse top is then clipped at a fixed (ideally) bias level and the two clipped pulses subtracted. The difference is the output of the phase detector. At the center position there is no output, hence the term balanced-phase detector. If the slope of the sawtooth is d, then the height of the output pulse is $2dx$ where x is the deviation of the pulse from its proper position. If the pulses are to the left of the center point, the polarity of the output (as well as any noise present) is reversed. Thus the output of this phase detector consists of pulses of current of varying amplitude and polarity. If, at time $t=0$, the frequency of the oscillator is correct but there exists a constant phase error, then the input to the servo consists of a sequence of pulses of the same height and polarity.

To make this type of input signal fit in with the previously derived equations describing the servomechanism, it is convenient to observe that the time constant of the over-all servo will be large compared to the time between pulses (of the order of 15 to 1) so that one may without appreciable error spread the area of the input pulse over the entire interval 0 to T. This provides a continuous input to the servo. The area of the error pulse is $2dxW$, where W is its width. The equivalent pulse after spreading then has a height $(2dxw/T)$. x is defined as the phase error. The quantity $2dW/T$ has the dimensions of amperes per unit phase error and is therefore defined as the sensitivity constant K_1 of the phase detector. Under these assumptions all of the preceding results on tracking, lock-in time, and so forth, are valid.

If the sync pulses are not gated, as is generally the case in black-and-white television, there is some difficulty in calculating the noise output of such a phase detector. To design for the worst condition, one would presumably assume the video at black level and then calculate the noise output of the device. This can be done only by numerical methods and will not be gone into here. In practice, the clipping levels of the two diodes in the balanced phase detector are allowed to set themselves. These levels are determined by the pulse signal level, the noise level, the pulse signal position, the diode internal resistance, and the diode plate load resistance. For a given input signal-to-noise ratio and with the input pulse at the neutral position, a solution can be found for the proper plate load resistor to position the clipping level where desired. For small errors in phase then the clipping level will not deviate much from the optimum value. The height of the sync pulse is $C/4$ where $(C/4) > 2de$ in all cases. Ideally, clipping should be done at $(C/4) - de$ since at this level full indication of phase error is obtained while the clipping level is as high as possible to avoid picking up noise. Presumably then, the clipping level is established for the minimum signal at which the device is expected to operate satisfactorily and suffers a design somewhat less than optimum at the higher signal-to-noise ratios. The plate load resistor thus determined affects the value of α in the filter following the phase detector but the capacity there can be varied within limits to adjust α for optimum value in terms of tracking, and so forth.

Appendix A

In the usual sync system, the sync tips are passed into the video and separated there by clipping. The top 25 per cent of the peak carrier may be devoted to hori-

[4] K. R. Wendt and G. L. Fredendall, "Automatic frequency and phase control of synchronization in television receivers," Proc. I.R.E., vol. 31, pp. 7–15; January, 1943.

zontal sync information for about 10 μseconds out of every line of 63.5 μseconds. The usable burst of carrier is probably 4 μseconds long and $C/4$ volts in amplitude where C is the intermediate-frequency carrier amplitude during sync pulse. It is assumed that this burst has been gated so that no noise appears in the sync system except during the burst itself. The noise in the intermediate frequency may be assumed to have a normal first probability density function of the form.

$$W_1(I) = \frac{1}{\sqrt{2\pi\psi_0}} e^{-I^2/2\psi_0}$$

where I is the instantaneous noise current and ψ_0 the average noise power. If the second detector is an envelope tracer, the density function of noise alone in the video is no longer normal but has a Rayleigh distribution

$$W_1(R) = \frac{R}{\psi_0} e^{\frac{-R^2}{2\psi_0}}$$

where R is the voltage of the envelope and ψ_0 the intermediate-frequency noise. If an unmodulated carrier C is added to the intermediate-frequency, the probability density function then becomes

$$W_1(R) = \frac{R}{\psi_0} e^{-(R^2+C^2)/2\psi_0} I_0\left(\frac{RC}{\psi_0}\right) \tag{1}$$

when I_0 is the modified Bessel function. This density function does not depend on the position of the carrier in the intermediate-frequency pass band. To a first approximation, the continuous part of the noise power spectrum $F(f)$ of a sine wave and noise in the video following an envelope is given by[5]

$$F(f) = \pi^2 h_{11}^2 [W(fq - f) + W(fq+f)] + \pi^2 \frac{h_{02}^2}{4} \int_{-\infty}^{\infty} W(x)W(f-x)\alpha x \tag{2}$$

where

$$h_{11} = \frac{1}{2}\left(\frac{y}{\pi}\right)^{1/2} {}_1F_1(1/2;\, 2;\, -y)$$

$$h_{02} = (2\pi\psi_0)^{-1/2} {}_1F_1(1/2;\, 1;\, -y)$$

$$y = \frac{C^2}{2\psi_0}.$$

${}_1F_1(\alpha;\, \beta,\, y)$ is the hypergeometric function. The first part of $F(f)$ is due to the modulation products of signal and noise while the second part is due to the intermodulation of the noise alone. In the case of vestigial side-band transmission, there is only half of the first part of this, namely, $W(fq+f)$ where now f extends over the entire intermediate-frequency and video bandwidth. If $W(f)$ is flat, $= W_0$, the noise power spectrum in the video is as shown in Fig. 5.

Here

$$A = (ch_{02})^2\left[\frac{BW_0^2}{4C^2}\right] = \frac{W_0}{8}(ch_{02})^2 \frac{1}{\left(\frac{S}{N}\right)_{\text{if}}}.$$

The values of $(ch_{02})^2$ and h_{11}^2 are obtained from the curves in the literature.[6] As the signal becomes larger compared with the noise, the contribution due to noise intermodulation drops off until ultimately the noise rides on top of the signal and the spectrum in the video has the same shape as in the intermediate frequency. On the other hand, for zero signal, the spectrum becomes triangular.

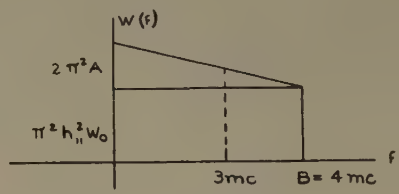

Fig. 5—Power spectrum of sine wave plus noise in the video.

The above shows the power spectrum of the noise alone. As is generally known, when the noise power becomes comparable to that of the signal, the second detector causes a suppression of modulation. Since the modulation index of the carrier burst is small (1,8), the preceding result concerning noise-power spectrum is not altered perceptibly by the addition of the modulation. However, it is necessary to know the extent of modulation suppression and it is convenient, since the modulation index is small, to use the procedure of Ragazzini.[7] The input to the detector can be written

$$e(t) = C \cos \omega_0 t + mC \cos (\omega_0 + \omega_s)t + \sum_0^{m_0} c_n \cos(\omega_n t + \phi_n)$$

where the last term encompasses all noise components with random phase angles ϕ_n. Then the envelope can be written as

$$E(t) \cong [C^2 + m^2C^2 + 2\psi_0 + 2mC^2 \cos \omega_s t]^{1/2}$$

where ψ_0 is the entire intermediate-frequency noise power. This is approximately

$$E(t) = \sqrt{C^2 + 2\psi_0}\left(1 + \frac{mC^2 \cos \omega_s t}{C^2 + 2\psi_0}\right).$$

[5] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 24, pp. 46–157; January, 1945. Rice's constant $1/\pi^2$ has been eliminated here.

[6] See page 148 of footnote reference 5.

[7] J. R. Ragazzini, "The effect of fluctuation voltages on the linear detector," Proc. I.R.E., vol. 30, pp. 277–287; June, 1942.

Thus the detected signal has the amplitude

$$\frac{mC^2}{\sqrt{C^2+2\psi_0}} = \frac{mC}{\sqrt{1+\frac{2\psi_0}{C^2}}}$$

and the modulation suppression is

$$\frac{1}{\sqrt{1+\frac{1}{\left(\frac{S}{N}\right)_{\text{if}}}}}.$$

For $S/N_{\text{if}}=1$, 3, and 5, the suppression factor becomes 0.71, 0.87, and 0.91.

Now, if in the video, the burst and associated noise is simply gated, it is possible, as is shown in appendix B, to find the signal-to-noise ratio of the 3-Mc sine wave and noise as it leaves the filter. However, some improvement can be made by clipping as well as gating. As shown in Appendix B, the noise spectrum at 3 Mc is, to a first approximation, reduced by the gating factor. The noise may be further reduced by clipping at $3/4C$ since this does not reduce the signal but does reduce the noise. Numerical integration of the second density function (1) shows that the total dc suppressed noise power is reduced from $0.65\psi_0$ to $0.50\psi_0$, $0.83\psi_0$ to $0.52\psi_0$, and $0.93\psi_0$ to $0.61\psi_0$, respectively, for input signal-to-noise ratios of 1, 3, and 5. Thus assuming that the spectral distribution does not change materially in shape, the power spectrum level at 3 Mc can be found.

Utilizing the results of (2), this is to be found to be $0.027W_0$, $0.032W_0$, and $0.037W_0$, respectively. Since ψ_0 the total intermediate-frequency noise $=BW_0$ this may be written

$$\frac{0.027\psi_0}{B}, \quad \frac{0.032\psi_0}{B}, \quad \text{and} \quad \frac{0.037\psi_0}{B}.$$

Appendix B

If we have an ideal gate which opens and closes at intervals T, it may be represented as the time function shown in Fig. 6. To gate noise this time function is

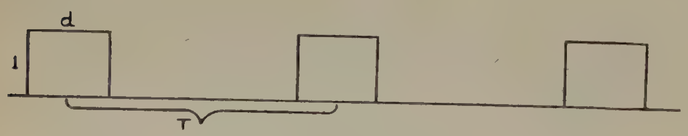

Fig. 6—Gating time function.

multiplied by the noise-voltage time function. It is desired then to find the power spectrum of the product. The autocorrelation of the product is the product of the separate autocorrelations since the two time functions are independent. To each autocorrelation function corresponds, by Fourier transform (Wiener's theorem), the power spectrum of the original time function, Thus,

$$G(\omega) = 4\int_0^\infty R(\tau)\cos\omega\tau d\tau$$

$$R(\tau) = \frac{1}{2\pi}\int_0^\infty G(\omega)\cos\omega\tau dw$$

where $G(\omega)$ is the power spectrum and $R(\tau)$ the unnormalized autocorrelation function. The spectrum of the product of two autocorrelation functions (time functions in the power domain) is given by the complex convolution integral. Thus the power spectrum of the two original time functions is given by

$$G(\omega) = \frac{1}{4\pi}\int_{-\infty}^{\infty} F_1(\omega-s)F_2(s)ds$$

where $F_1(S)$ is the power spectrum of the noise, say, and $F_2(S)$ is the power spectrum of the gating pulses. For the simple gate assumed, the result is given by

$$F(\omega) = \frac{d^2}{T^2}\sum_{-\infty}^{\infty}\left[\frac{\sin\frac{n\pi d}{T}}{\frac{n\pi d}{T}}\right]^2 G_1\left(\omega-\frac{2\pi n}{T}\right).$$

This sum may be approximated by an integral

$$F(\omega) \cong \frac{d^2}{T^2}\int_{-\infty}^{\infty}\left[\frac{\sin\frac{n\pi d}{T}}{\frac{n\pi d}{T}}\right]^2 G_1\left(\omega-\frac{2\pi n}{T}\right)dn$$

which is valid when the gating frequency is small compared to the bandwidth of the noise, so that successive G's overlap substantially.

If $G_1(\omega)$ is constant and $=G_{11}$

$$F_1(\omega) = \frac{G_{11}d^2}{T^2}\left[1+2\sum_1^\infty\left[\frac{\sin\frac{n\pi d}{T}}{\frac{n\pi d}{T}}\right]^2\right] = \frac{G_{11}d}{T}$$

showing that the noise spectrum and the total noise power are each reduced by the gating factor d/T. By the same argument one may show that for any spectrum $G_1(\omega)$, the total noise power is reduced by d/T.

It may be of some interest to observe here that when gates are used in the sense of samplers of a video signal, the signal-to-noise ratio after gating does not depend on the gating factor d/T in any way, provided that the noise bandwidth is limited to ω_0 prior to sampling where ω_0 is the highest video frequency sampled. This is apparently not generally realized but may be easily shown with the aid of the above formulas.

In the application used here, the gating frequency is small compared with a bandwidth of 4 Mc so that the integral approximation is valid. Under these conditions, it is not hard to show that to a first approximation the

spectrum at 3 Mc is also reduced by d/T when the total power is reduced by that amount.

Appendix C

In the case of normally distributed noise in a range of frequencies that is small compared to the center frequency, the noise current may be written after the method of Rice[8] as

$$I_n = \sum_1^N C_n \cos(\omega_n t + \phi_n)$$
$$= \sum_1^N C_n \cos[(\omega_n - q)t + \phi_n + qt]$$
$$= I_c \cos qt - I_s \sin qt$$

where

$$I_c = \sum_1^N C_n \cos[(\omega_n - q)t + \phi_n]$$
$$I_c = \sum_1^N C_n \sin[(\omega_n - q)t + \phi n].$$

Here q is the center frequency of the band,

$$\omega_n = 2\pi f_n, \quad f_n = n\Delta f, \quad C_n{}^2 = 2F(f_n)\Delta f.$$

$F(f)$ is the noise power spectrum. I_c and I_s are normally distributed and each have the rms value $\sqrt{\psi_0}$. If a sinusoid $Q \cos qt$ is added to the noise,

$$I = Q \cos qt + I_n$$
$$= (Q + I_c) \cos at - I_s \sin qt$$
$$= R \cos(qt + \theta)$$

where R is the envelope and θ the phase. Then

$$\theta = \tan - 1 \frac{I_s}{I_c + Q} \cong \frac{I_s}{Q} \quad \text{if} \quad Q \gg \sqrt{\psi_0}$$

and under these conditions the density function for θ may immediately be written as

$$W_1(\theta) = \frac{Q}{\sqrt{2\pi\psi_0}} e^{\frac{-\theta^2 Q^2}{2\psi_0}}.$$

Then the rms phase deviation from that of the sinusoid is $\sqrt{\psi_0}/Q$ radians. From the representation of noise in the Fourier series, it is evident that I_s behaves like a noise current whose power spectrum is concentrated in the power part of the power spectrum and is, in fact,

$$F_1(\omega) = F(\omega_q + \omega) + F(w_q - \omega)$$

where $F(\omega)$ is the power spectrum of I_n. Thus if

$$F(\omega) = \begin{cases} F_0 = \dfrac{\psi_0}{B} & \text{for} \quad \omega_q - \dfrac{\beta}{2} < \omega < \omega_q + \dfrac{\beta}{2} \\ 0 & \text{elsewhere,} \end{cases}$$

then

$$F_1(\omega) = \begin{cases} \dfrac{2F_0}{Q^2}, & 0 < \omega < \dfrac{\beta}{2} \\ 0 & \text{elsewhere.} \end{cases}$$

The noise in the video is not normally distributed. However, if a narrow filter at 3 Mc is inserted, randomness is restored and the noise out of the filter can be considered normal, so that the preceding argument can be applied. The filter may be assumed so narrow that the noise spectrum picked out by it will be flat. However, the level of the noise power at 3 Mc must be determined.

Using the results of Appendix B, it can be shown without difficulty that the condition $Q \gg \sqrt{\psi_0}$ is satisfied for the three values of S/N_{if} assumed. A filter 2 kc wide will quite satisfactorily pick out the 3-Mc carrier without the adjacent sidebands which are 15 kc away. This filter width does not give rise to any appreciable static phase error in itself due to changes in frequency and provides a Q suitably larger than $\sqrt{\psi_0}$.

It is of interest to note here that the phase noise-power spectrum has the dimensions of time so that when integrated over ω, it yields phase in radians. The phase correlation may also be easily calculated from the phase power spectrum, being given by Wiener's theorem as

$$R(\tau) = \frac{1}{2\pi} \int_0^\infty F(\omega) \cos \omega\tau dw.$$

In the case of a uniform spectrum of width γ this yields

$$R(\tau) = \frac{F_0\gamma}{4\pi} \frac{\sin \frac{\gamma\tau}{2}}{\frac{\gamma\tau}{2}}$$

showing that phase correlation due to noise does not become small until

$$\tau = \frac{2\pi}{\gamma}.$$

[8] S. O. Rice, "Statistical properties of a sine wave plus random noise," *Bell Sys. Tech. Jour.*, vol. 27, pp. 109–158; January, 1948.

The Control Chart as a Tool for Analyzing Experimental Data*

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This paper is published at the suggestion and on the recommendation of the IRE Professional Group on Quality Control.

Summary—**The statistical methods that have been developed for use in quality control are a powerful tool in the interpretation of laboratory experiments where only a small amount of data is available. An understanding of these methods also permits more logical planning of experiments and improves what we might call "the efficiency of experimentation." One of the simplest and most broadly useful of these tools is the control chart. It is easy to understand and use and in many cases can take the place of more laborious and complicated methods of analysis.**

ONE OF THE simplest and most useful tools for statistical analysis is the control chart. As originally developed by Shewhart, the control chart was primarily a tool for the production man who dealt with large numbers of repetitive operations. In recent years, Shewhart and others have been giving much thought to the question: "Can the simple statistical tools that have proved so valuable in quality control in the factory be just as useful in research and development work, in laboratory experimentation?" The answer is yes.

It is the purpose of this paper to give a very brief description of the control chart and point out how the experimenter can use it in his laboratory. He may not be an expert statistician. But he wants to make most of the analysis by himself and call in the expert only on the more difficult problems. He needs a technique that is not cloaked in mathematical symbolism. He needs a method of analysis that can be applied in a straightforward manner, and that can be explained to others who, like himself, are not steeped in statistical lore. He is more interested in relations and interpretations that lead to sound and useful judgments than he is in mathematical elegance or numerical precision.

The control chart pays its way as a simple and efficient method of analyzing and summarizing our data. Then it pays a bonus. It tells us when our data are good and when not. It tells us when we have enough data and when we need more. It tells us when our methods can be improved and when we have reached the limit in refining them along any particular line. It points a finger at those spots in our development where concentrated effort will do the most good.

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† Bell Telephone Laboratories, Inc., New York, N. Y.

In appearance, the control chart is a simple plotting of points and guide lines. Typically it is a double chart like that in Fig. 1. The abscissa is some independent variable such as observer, the number of the subgroup or setup, or ambient conditions. Often it is simply a serial number representing the order in which the experiment is carried out. The ordinate is the characteristic of interest: some physical property of an alloy under study, or the transconductance of a tube of new design.

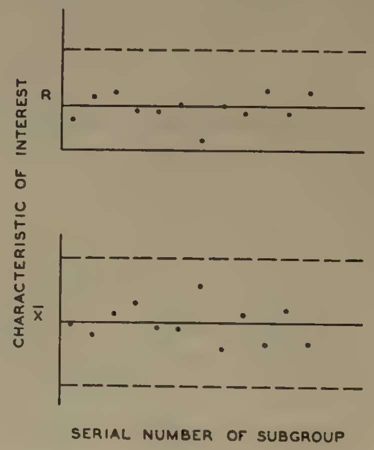

Fig. 1—Control chart subgroups of 4.

In plotting our control chart, we group our data into small rational subgroups. The "small" means, preferably, 4 or 5, although we sometimes use subgroups as small as 2 or as large as 6 or more. The "rational" means that we have reason to believe the members of a subgroup are alike because they involve the same calibration setting, or the same bottle of reagent, or simply because they represent work all done at about the same time. The "rational" also means that we have reason to suspect that the subgroups might be different from each other, because they represent different observers, or different calibrations, or simply different days.

We treat the few observations in each subgroup as a set and we compute and plot their range R and their average $\overline{X}$. The range is the difference between the largest and smallest, and is a measure of dispersion. The average is the usual arithmetic mean. The range and the average are the points on the control chart.

On our control chart we draw two central lines, one

at the average of the ranges $\overline{R}$; one at the average of the averages, which is the grand average $\overline{\overline{X}}$.

On the control chart we also draw limit lines.

$$\lim R = D_3\overline{R} \quad \text{or} \quad D_4\overline{R}$$
$$\lim \overline{X} = \overline{\overline{X}} \pm A_2\overline{R}$$

where D_3, D_4, and A_2 are factors that we find in published tables,[1] just as we find sines or logarithms. They depend on the number of observations in each subgroup.

The central line on the average chart represents the average in the usual sense. It is the average strength of our plastic, the average permeability of our iron, or the average g_{mi} of our tubes.

The central line on our range chart tells us what dispersion to expect. If these data represent reliable information about a stable process, then we expect future observations to have a root-mean-square deviation σ of about

$$\sigma = \overline{R}/d_2$$

where d_2 is another factor from the tables,[1] and depends on the subgroups size. We expect nearly all future observations to lie between $\overline{\overline{X}}+3\sigma$ and $\overline{\overline{X}}-3\sigma$.

The limit lines tell us something about the reliability and stability. If we have points outside limits we should question the reliability and stability. Points outside, outages, tell us that our experimental process is probably being affected by differences in calibration, by the mood of the observer, or by the intrusion of some foreign influence. Outages tell us that our measurements are being affected by some identifiable or assignable cause for variation—a cause not normally present—a cause that stands out above that system of causes that produce the variations inherent in our process. Outages tell us that our experiment is in trouble and tell us when it got that way. This information is most important in tracking down and eliminating that trouble.

If there are no outages, this indicates that the variations we observe are due to a mass of minor causes, and not to any one or two causes that are important or even identifiable. So we know that we should not waste time tinkering with the process in an effort at improvement. For this indication to be reliable, we should have something like 25 consecutive subgroups all within limits.

As a corollary to this, the limit lines sometimes tell us when to abandon a particular process or type of experimentation. The limits which the process gives us for itself way be wider than we can tolerate. But if we have enough points all inside limits, with no outage, and still want something better, we must make a fundamental change in the process.

We should emphasize the fact that these limits are not based on any customer's requirements, or on management's stated objectives, or on what some old hand tells us we ought to be able to do. They are based on the measurements we have just made on the process we are working with. By these limits the process itself tells us what to expect from it in the future.

The control chart is a system of bookkeeping for studying the causes of variation. We use it to analyze the same data we would analyze by any other method. It helps us draw the same conclusions we ought to draw by whatever method we use. It is simple and straight forward.

Let us consider a couple of simple illustrations of how the control chart might be useful in laboratory work. Suppose we have a paper design for a new type of vacuum tube. We make a small batch of tubes according to this design—4 or 5, or maybe 8 or 10. We measure their characteristics. For this illustration, let us confine our attention to the trans-conductance g_m. Perhaps our first models do not turn out as well as we had hoped. We revise our design and make another lot. It is still pretty poor, but we think we are beginning to see what is wrong so we try again. This time we get something worth while. This is the familiar history of experimental progress. Finally, after more trials, we arrive at a design that we think may be satisfactory. We take it to the production man and the customer for their opinions.

They like it, and the eight tubes we made according to the last design look good. Then comes the inevitable question, shot at us in two forms. From the user: "How much variation will I have to put up with?" From the producer: "How tight tolerances must I meet?"

Fig. 2 shows a simple historical record of all the tubes we have made. In preparing this hypothetical example, it was assumed that nine different designs, A, B, $\cdots$ J

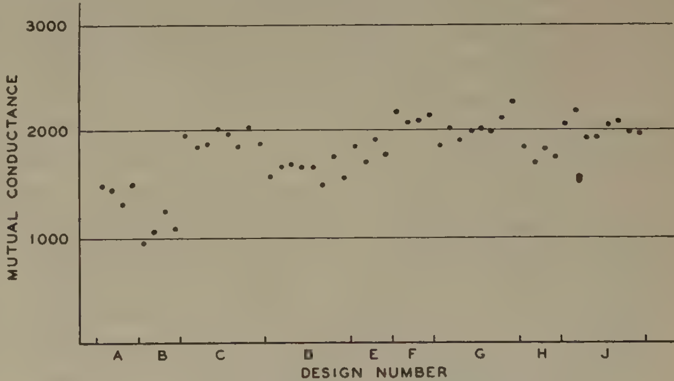

Fig. 2—Historical record. Individual points.

on the chart, were tried and that the later ones were the better ones. The g_m varies all the way from 950 to 2,270. It is not apparent from this chart just what variation is to be expected.

Now let us put these same data in control-chart form as shown in Fig. 3. Some of our design lots contained four tubes. Others contained eight and we have treated such a lot as two subgroups of four. Altogether, we have thirteen subgroups of four. Design changes have been made between subgroups so that any subgroup relates to a single design.

[1] "A.S.T.M. Manual on Presentation of Data," American Society for Testing Materials, Philadelphia, Pa.; April, 1945.

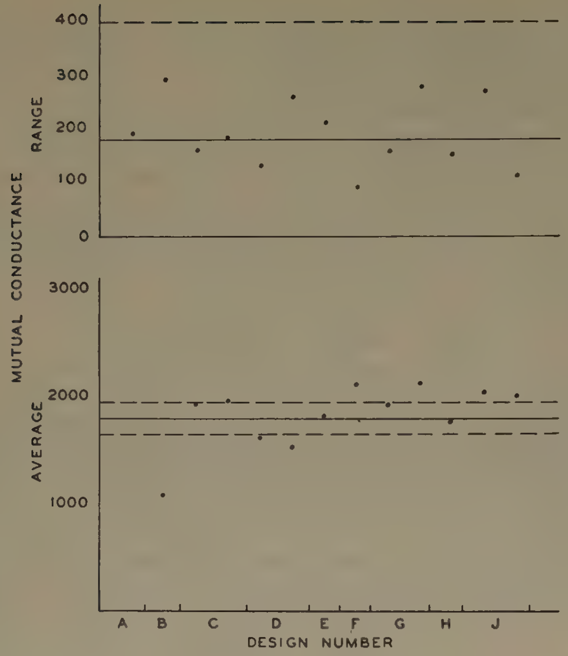

Fig. 3—Historical record in control-chart form.

From the control chart of averages we see, as was assumed in the illustration, that designs *A*, *B*, and *D* give significantly low g_m while designs *C*, *F*, *G*, and *J* give significantly high g_m.

These control charts should, of course, have been plotted day by day as our work progressed. In the early stages, we would have used tentative values, based on the data available then, for the central lines and limits. These control charts should have been used to guide each step in our development. They would have helped us to evaluate the intended effects of our design changes. They would have helped us recognize unintended effects which we might desire either to avoid or to use. They would have helped us to recognize and eliminate extraneous effects that might interfere with our experimentation.

The range control chart does its share in this guiding of the course of our work. It also helps us answer that question about variations or tolerances. Every time we changed our design we may have changed our design average. If we computed a single dispersion on the basis of all the tubes lumped together, that dispersion would probably describe principally our design changes. But within each subgroup of four there was no design change. The range of each subgroup represents the dispersion that is present with constant design. We have thirteen such ranges, and from their average we can form an estimate of the inherent dispersion that exists when the design is not changed, and that is based on an over-all sample of 52 tubes and not merely on the eight tubes of the final design.

One may well ask: But may not these intentional design changes have produced changes in inherent dispersion just as they did in the average? The answer obviously is: Yes, they may have. Fortunately, it is rather common experience that they do not. But let us not take this for granted. Let us go back to the limit line and the outages on our range chart. In this illustration it was assumed that there were not outages. This is an indication that there was no significant change in dispersion, due either to the design changes or to extraneous causes. We may, therefore, take the average range as a measure of dispersion with some confidence.

But suppose that for some subgroups the range is outside limits. This may indicate that this design tends to give greater inherent variation than the others and may cause us to abandon this design. Or it may indicate the intrusion of some foreign influence into our model building or our measurements. In either case, we would not include the range of this subgroup in the basis of our estimate of dispersion.

We may compare this method to the analysis of variance. The latter also gives us a measure of the changes that our designs have introduced and a measure of the inherent or residual dispersion that exists with a fixed design. The analysis of variance, however, makes no use of limits and outages. The analysis of variance assumes that the residual variation is constant. It waves no red flags when the inherent dispersion changes or when extraneous influences interfere. It does not tell us to ignore certain subgroups in estimating the residual dispersion.

In planning this series of experiments we would have used our knowledge of our intended method of analysis to plan the size of our subgroups, to plan the essential similarity of items within any one subgroup, and to plan specific changes between subgroups.

Now consider an example in which statistics play a more important part in planning an experiment. Suppose we are working on a 3-stage intermediate-frequency amplifier for a television receiver and want to determine experimentally the variation in gain that is to be expected from variation in certain components. This gain depends, among other things, on the g_m's of the three tubes. We might get the three worst tubes we we can find, and the three best tubes we can find, and try each set in our circuit. This is unrealistic and may be unnecessarily pessimistic. From our knowledge of of the dispersion of this type of tube, and by the use of known statistical factors[2] we can write down the g_m for each tube in an idealized sample of nine. We know, for example, that, on the average, the largest value in a sample of nine from a normal universe will be 1.48 σ above the universe average. Let us get a number of tubes from the stockroom, measure each, and select the nine which most nearly correspond to the idealized sample. We also know the expected values of the three averages we would get by an idealized sampling process which gave us three sets of three tubes each. So we divide our nine tubes into three sets of three each whose averages most nearly correspond to those expected

[2] Cecil Hastings, Jr., F. Mosteller, J. W. Tukey, and C. P. Winsor, "Low moments for small samples," *Inst. Math. Stat.*, vol. 18, no. 3; September, 1947.

averages. This is illustrated in Fig. 4. We put each of these three sets of tubes in our circuit and measure the resulting gain.

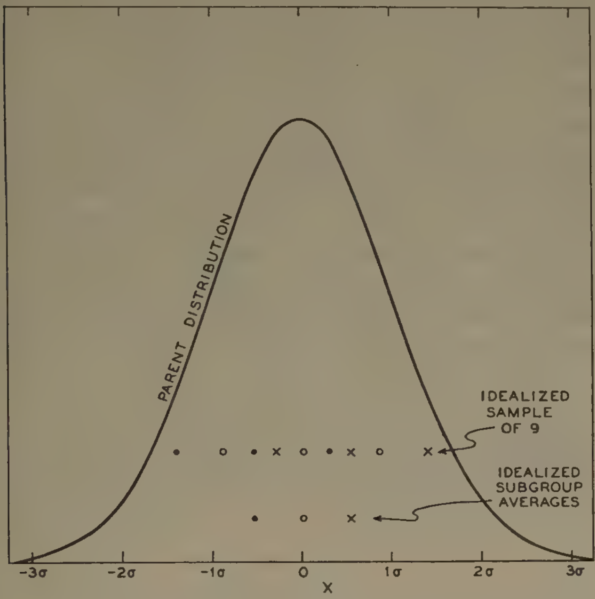

Fig. 4—An idealized sample of 9 items selected according to some important parameter and arranged in subgroups of 3.

There are probably some other circuit elements that are important to the gain of the amplifier—perhaps the three biasing resistors. Here again, we can pick three similarly selected sets of biasing resistors to form purposeful samples just as we did with the tubes. We would build three amplifier circuits, one with each set of resistors. Then we would test each set of tubes in each circuit.

We might make each test at three supply voltages: rated voltage, 10-per cent high and 10-per cent low. This gives us a total of 27 gain measurements to be made.

This gives us no guarantee against errors in judgment as to the most important variables nor against the intrusion of other factors into our experiment. But this kind of logical planning will increase the amount of information and the reliability of the information that we get from a given amount of effort and equipment.

Note that this planning assumed that we knew the dispersion of the tubes and of the resistors. This means that the supplier of our components has been keeping control charts of his product and that he can and will give us his pertinent quality data. With this co-operation from our supplier we can make enormous improvements in the efficiency of our experimentation.

After we make the measurements we must analyze the data. We might do this in the conventional manner by the analysis of variance. Our assumed experiment involves a three-way classification according to transconductance, cathode resistor, and supply voltage. We have no explicit replication, that is, no simple repetition, but each measurement is affected by the residual dispersion that we would observe if we made simple repetitive observations under each condition. The analysis of variance gives us quantitative measures of the three main effects. It tells us, for example, how much variation in gain was related to the variation in transconductance. It also gives us quantitative measures of the interactions, such as the effect that variation in biasing resistors had on the dispersion due to variation in transconductance. And it gives us a quantitative measure of the residual dispersion.

We can get comparable information from an analysis by the control-chart method. We can get it more readily, and we can get it in a form that the ordinary engineer can understand. Let us outline a schedule for such an analysis. Part of such a schedule is shown in Table I. Columns a, b, and c give the values of the three independent parameters, and these are arranged in some logical order such as that indicated. In our example these are transconductance and biasing resistors, in each case the average of the three used, together with the supply

TABLE I

Analysis of Variation by Control-Chart Methods

a	b	c	d	e	f	g	h	j	k
(g_m)	(R_K)	(E_B)	(Gain)	R_c	$\bar{X}_c$	R_b^*	$\bar{\bar{X}}_b^*$	R_a^*	$\bar{\bar{X}}_a^*$
1	1	1	—						
		2	—						
		3	—	—	—				
	2	1	—						
		2	—						
		3	—	—	—				
	3	1	—						
		2	—						
		3	—	—	—	—	—		
2	1	1	—						
		2	—						
		3	—	—	—				
	2	1	—						
		2	—						
		3	—	—	—				
	3	1	—						
		2	—						
		3	—	—	—	—	—		
3	1	1	—						
		2	—						
		3	—	—	—			—	—
	2	1	—						
		2	—						
		3	—	—	—			—	—
	3	1	—						
		2	—						
		3							
Average			$\bar{\bar{X}}$	$\bar{R}_c$	$\bar{\bar{X}}$	$\bar{R}_b^*$	$\bar{\bar{\bar{X}}}$	$\bar{R}_a^*$	$\bar{\bar{\bar{X}}}$
Limits			{	—	—	—	—	—	—
				—	—	—	—	—	—

Note—a, b, and c are independent parameters. X is quantity measured. R_c is range of subgroup in which c is only variable parameter. $\bar{X}_c$ is similar average. R_b^* is range of a subgroup of $\bar{X}_c$ in which b is only variable parameter. Similarly $\bar{X}_b^*$, $\bar{R}_a^*$, and $\bar{X}_a^*$. σ_a, σ_b, and σ_c are estimates of main effect dispersions. σ_r is estimate of residual dispersion. d_2 and factors for computing limits are taken from published tables.

$$\sqrt{\sigma_c^2 + \sigma_r^2} = \bar{R}_c/d_2 \tag{1}$$

$$\sqrt{\sigma_b^2 + (\sigma_c^2 + \sigma_r^2)/3} = \bar{R}_b^*/d_2 \tag{2}$$

$$\sqrt{\sigma_a^2 + (\sigma_c^2 + \sigma_r^2)/3} = \bar{R}_a^*/d_2. \tag{3}$$

Rearrange primary data similarly in orders b, c, a and c, a, b. Obtain two solutions each for σ_a, σ_b, and σ_c. Average each pair. Obtain three solutions for σ_r. Average. Significant interactions and unstable residuals appear as ranges outside limits. Significant main differences appear as averages outside limits.

voltage. In the table we have called the three values of each parameter simply 1, 2, and 3. We might have called them g_{m1}, g_{m2}, $\cdots E_3$. In column d is the observed quantity in which we are interested, the gain. These 27 measurements, in the order shown, fall into nine subgroups within each of which voltage is the only variable parameter. These are rational subgroups in so far as the transconductance and biasing resistor are constant within groups but variable between groups.

In column e are the ranges of these subgroups. They represent a combination of two dispersions, one the inherent variation that would be present if we attempted simple repetitive measurements, the other the variation that is due to the variation in voltage. This is stated in (1) of Table I. We have called these ranges R_c since column c contains the only variable parameter in the subgroup.

We also compute the averages of each of the nine subgroups, as shown in column f. In these averages $\overline{X}_c$, the effects of the two causes just mentioned have been reduced by the averaging process. We can divide these averages $\overline{X}_c$ into secondary groups so that g_m is constant within any secondary group but variable between secondary groups. The ranges of these secondary groups R_b^* are shown in column g. The average of these ranges represents the dispersion due to biasing resistors, in combination with the dispersion already measured and here reduced by the first averaging process. This is shown in (2). Equations (1) and (2) may now be easily solved for the dispersion due to resistors alone.

By carrying out this last process with a different grouping of our first averages, $\overline{X}_c$ in column f, so that the resistors are constant within each new secondary group, we find the dispersion due to g_m alone. This is indicated in column j and (3). The ranges used here are called R_a^*.

Then we go back to our original data and rearrange it so tubes are the only variables in our first subgroups. We repeat the analysis as outlined above and obtain the dispersions due to resistors and voltage. Similarly, with another rearrangement of the original data, we obtain the dispersions due to voltage and tubes.

We now have two determinations or, more properly, two estimates of each main effect σ_a, σ_b, and σ_c. Because of sampling fluctuation, and because we have ignored the interactions, these two determinations may not be in perfect agreement with each other or with that made by an analysis of variance. Some experience has indicated that the average of these two determinations is a useful estimate of the main effect.

Finally, we compute the residual dispersion. We had determined it in combination with each main effect. We now know these main effects, we subtract them out and average the three remaining determinations to obtain an estimate of the residual dispersion.

So far we have ignored the interactions. We have no numerical values for them, such as the analysis of variance gives. But at each stage of our analysis we will have computed control-chart limits for all the subgroup and secondary ranges and averages mentioned, and for the secondary group averages, $\overline{X}_b^*$ in column h and $\overline{X}_a^*$ in column k. Outages in our ranges tell us which interactions are significant just as outages in our averages tell us which main effects are significant.

These interactions may be shown by a mass chart of outages, as in Table II. Here, in Part A, are all the subgroup ranges R_c of column e. This table is a rectangular array with the ranges for one set of tubes in one column and the ranges for another set of tubes in another column. That puts the ranges for one set of resistors in one row and the ranges for another set of resistors in another row. Part B is a similar table for the averages of column f.

TABLE II

MASS CHART OF OUTAGES

(From an experiment similar to that outlined in Table I.)

A. Ranges for 7 Voltages[1]

<table>
<tr><th></th><th>Tubes</th><th>1</th><th>2</th><th>3</th><th>4</th><th>5</th><th>6</th><th>7</th><th>8</th><th>9</th><th>10</th><th>11</th><th>$\overline{R}_c$</th></tr>
<tr><td>Resistors</td><td>1</td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td>|</td><td></td></tr>
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<tr><td></td><td>7</td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td></tr>
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<tr><td></td><td>$\overline{R}_c$</td><td></td><td></td><td></td><td></td><td></td><td>—</td><td>—</td><td></td><td></td><td></td><td>|</td><td></td></tr>
</table>

B. Averages for 7 Voltages[1]

<table>
<tr><th></th><th>Tubes</th><th>1</th><th>2</th><th>3</th><th>4</th><th>5</th><th>6</th><th>7</th><th>8</th><th>9</th><th>10</th><th>11</th><th>$R_b^*\overline{X}_b^*$.</th></tr>
<tr><td>Resistors</td><td>1</td><td>—</td><td>—</td><td>—</td><td>—</td><td></td><td></td><td></td><td></td><td></td><td>|</td><td>|</td><td></td></tr>
<tr><td></td><td>2</td><td>—</td><td>—</td><td>—</td><td>—</td><td></td><td>—</td><td></td><td>—</td><td></td><td>|</td><td>|</td><td></td></tr>
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<tr><td></td><td>R_a^*</td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td></td><td>|</td><td></td></tr>
<tr><td></td><td>$\overline{X}_a^*$</td><td>—</td><td>—</td><td>—</td><td>—</td><td>—</td><td>—</td><td></td><td></td><td></td><td>|</td><td>|</td><td></td></tr>
</table>

[1] | = Values above upper limits. — = Values below lower limits.

Now this is to be a study of outages. So, on these charts, we write no numerical values. If a range or average is within limits we leave a blank. It is outside limits and high, we write a short vertical bar. If it is an outage and low, we write a short horizontal bar. On work sheets, we can use blue circles for high values and red circles for low values.

The data for Table II were taken from an experiment that had nothing to do with amplifiers. It has been relabeled to fit the hypothetical example of Table I. There were eleven sets of tubes, ten sets of resistors, and measurements were made at seven different voltages.

Part A shows a large number of range outrages for

tubes 11. This means either that the tubes in this set are more sensitive to voltage change than the others, or that they have a higher residual dispersion. In either case, this is a significant interaction. It is indicated by a pattern of outages.

In the range table is a single outage under tubes 5. This is a "wild shot."

In Part B, showing outages of averages, we see that the higher-numbered tubes give higher gain and that the lower-numbered resistors give lower gain. These are main effects.

It has been mentioned that data of this type could be studied by the analysis of variance. The control-chart method has its advantages. For one thing, it requires less work. Some experience indicates that the control chart method takes less than half the time required for an analysis of variance.

More important is the ability of the control chart to pinpoint troubles and to test the validity of the assumptions made.

A large number of outages on our control charts, arranged in some logical pattern, gives us good indication of both the existence and the nature of real differences or real interactions. A fair number of outages, scattered over the whole experiment in an irregular fashion, usually indicates poor experimentation—sloppy work. A small number of outages, particularly if they can be traced back to a few pieces of primary data, indicates that these observations were abnormal—"wild shots."

The analysis of variance assumes a stable residual dispersion and then forgets it. The control chart sets a trap for instability and waves a red flag when it appears. The analysis of variance assumes that certain interactions are possible and ignores all others. The control chart method makes an initial assumption of no interaction and then highlights any interaction that comes along.

The control chart continually invites us to examine the quality of our experimentation and furnishes us with the means for making that examination.

Statistical Evaluation of Life Expectancy of Vacuum Tubes Designed for Long-Life Operation*

ELEANOR M. McELWEE†

Summary—**Life-test data on subminiature vacuum tubes designed for 5,000 hours are analyzed statistically and an equation is derived for the curve of life survival percentages. Correlation of individual types to the general curve is found to be extremely high. Controls are determined for normal 500-hour life tests which assure rated long-life quality and are presented as a method for evaluating life expectancy before completion of long-life tests. Life test samples of lots of tubes released by this 500-hour plan were continued in operation for 5,000 hours, and the results are shown to be satisfacfactory.**

IN RECENT YEARS, the expanding field of industrial applications of vacuum tubes has contributed to an increased demand for greater reliability over a longer period of time. In reponse to this demand, engineers throughout the industry have attempted to design a line of electron tubes which might safely be rated far beyond the customary 500 hours used to evaluate the life of radio receiving tubes. The innovations in design and processing which produced the longer-life tubes, although undoubtedly of tremendous interest, are not within the scope of this paper. The problem with which we are concerned is one introduced by the development of such tubes—that of evaluating "long-life" quality within a reasonable length of time. It is obviously impractical for tube manufacturers to conduct life tests for 5,000 hours before release of lots of tubes, or for customers to wait seven to eight months for delivery. The efforts of many engineers were therefore applied to the search for a life test plan which would effectively measure 5,000-hour quality within the normal 500-hour life test period.

The life of a vacuum tube is commonly understood to be the length of time it will operate within a specified range of characteristics. Thus a tube is considered to have reached the end of life when it becomes inoperable for any reason, or when its characteristics fall outside the end point limits specified for the particular type. The normal life rating of 500 hours did not guarantee, however, that each tube had a minimum life of 500 hours, nor even that the average life of a group of tubes would be 500 hours. As defined in the JAN-1A specification, a 500-hour rating guarantees an aggregate useful life of at least 80 per cent of the total rated life. For example, a group of five tubes would have a total rated life of 2,500 tube hours. These tubes would pass the JAN specification if their total operation within specified limits was 2,000 tube hours or better. This total figure could be amassed in any one of a number of ways; e.g., by all tubes operating for 400 hours, by one failure immediately and four good to 500 hours, by two tubes good to 250 hours and three good to 500 hours, and so on. What is actually required is an average life of at least 80 per cent of the rating for any group of tubes life-tested for

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† Sylvania Electric Products Inc., Kew Gardens, L. I., N. Y.

the specified time. For the purpose of better understanding of this paper, it will be assumed that a 5,000-hour life rating is applied in the same manner; i.e., that the 80 per cent limit will apply to a 5,000-hour, rather than a 500-hour test point, and that the average life of any group of tubes must be at least 4,000, rather than 400 hours.

It was apparent in the beginning that there were two approaches to the problem of a shorter life test: (1) an accelerated test which would be equivalent to 5,000 hours of normal operation, or (2) statistical controls on a normal 500-hour life test which would adequately predict 5,000-hour quality. Considerable time and effort were expended in the attempt to set up a reliable accelerated test. Various changes were made in voltages, currents and/or power dissipations in the hope of discovering a test exactly ten times as rigorous as normal operation. Unfortunately, it was impossible to determine test conditions which would accelerate normal tube failures without introducing contributory factors not present in normal operation. Several tests were found satisfactory for individual types of failures; e.g., cycling tests to determine the quality of heaters, immersion test for air leaks, fatigue test for shorts or poor welds, etc. However, it remained impossible to obtain satisfactory correlation between emission deterioration resulting from any accelerated test, and that resulting from normal operation. Consequently, the emphasis was transferred to statistical analysis in the hope of determining a consistent pattern to which the quality control method could be applied.

The first step in the statistical analysis of data was logically a survey of the occurrence of failures in operation with relation to time. In order to include results on as many tubes as possible, the initial survey was made on life test samples of early subminiature indirectly heated cathode-type vacuum tubes, rated for normal 500-hour operation. Failures per 500-hour period were listed for a heterogeneous group of 1,864 vacuum tubes, and the average life percentage[1] at the end of each period was calculated in accordance with the JAN specification, as shown in Table I. The ratio between these percentages seemed to indicate that they would follow the exponential curve $y=ab^x$, where y = average life percentage, x = hours of life expressed in thousands of hours, and a and b are constants denoting the y intercept and the slope of the line, respectively. In order to determine the goodness of fit of the empirical curve $y=ab^x$, or the straight line $\log y = \log a + x \log b$, the values of a and b were found by the statistical method of least squares.[2] Then the calculated equation becomes

$$y = 93.1(0.875)^x.$$

By substituting the given values of x, computed values of y are obtained, and y residuals are found by subtraction. From the statistical formula for the standard error of estimate,

$$S_y = \left[\frac{\sum (y \text{ observed} - y \text{ computed})^2}{\text{the number of observations}}\right]^{1/2}.$$

$$S_y = (4.05406)^{1/2}.$$

The index of correlation of the curve is determined by the formula

$$\rho_{xy} = \left(1 - \frac{S_y^2}{\sigma_y^2}\right)^{1/2},$$

where S_y is the standard error of estimate of the curve, and σ_y is the standard deviation of the observed values of y. Substituting,

$$\rho_{xy} = \left(1 - \frac{4.05406}{173.333}\right)^{1/2}$$

$$\rho_{xy} = 0.988.$$

The high degree of correlation obtained was a positive indication that the empirical equation $y=ab^x$ was a close representation of these data, at least. In order to verify the results of this first experiment, the same method was followed with two additional groups of data. For a group of 1,240 vacuum tubes of various types, most of which were experimental tubes designed for a longer life rating, the calculated curve was $y=98.2(0.966)^x$. The index of correlation with observed data was 0.998. For a group of 130 tubes of six types released as 5,000-hour tubes, the equation of the calculated curve was $y=98.3\ (0.973)^x$; the index of correlation was 0.995. Both the observed data and the calculated curve are plotted for each group in Fig. 1.

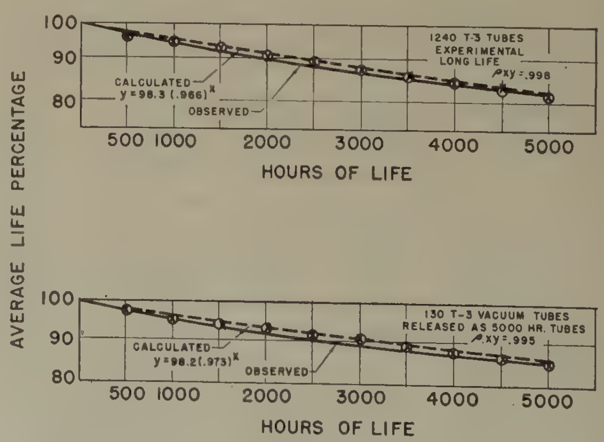

Fig. 1—Life survival curves: $y=ab^x$. Observed data and calculated curves for two heterogeneous groups of longer-life tubes.

[1] Average life percentage at X hours =

$$\frac{\Sigma \text{ (life hours for each tube*)}}{X \text{ hours (number of tubes started)}} \times 100\cdot$$

E.g., if 5 tubes were started on life, one failed at 700 hours, 4 remained good past 1,000 hours, the average life percentage at 1,000 hours would be

$$\frac{700 + 4(1000)}{5(1000)} \times 100 = 94 \text{ per cent.}$$

* The life for any individual tube shall be a maximum of X hours.

[2] C. H. Richardson, "An Introduction to Statistical Analysis," Revised Edition, Harcourt, Brace and Co., New York, N. Y., pp. 210–219; 1944.

With the acceptance of the curve $y=ab^x$ as a general pattern for life survival percentages, there remained two essential points to be determined: (1) Could a universal value of the constant a be assumed that would satisfy all types of tubes? (2) If the value of the constant b were calculated from observed 500-hour results, how closely would the predicted percentages approximate actual life test operation? The answer to the former question at first seemed evident. Since only good tubes are subjected to life test, it was assumed that the y intercept of the curve would be 100 per cent, and therefore the equation would be $y=100b^x$. Accordingly, the equation was checked with observed data, but it was noted that actual life test results beyond 1,500 or 2,000 hours were in all cases better than predicted percentages. Further analysis of data revealed that the rate of failure during the first 500-hour period of operation was higher than the rate of failure for any succeeding 500-hour period. The data seemed to indicate, in fact, that the rate of failure beyond 500 hours would be fairly constant, and would be approximately half that of the first 500-hour period. To compensate for this phenomenon, it was decided to use the value 99 for the constant a. In order to check the validity of predicted percentages, the same groups of data used previously were checked with percentages calculated from the equation $y=99b^x$, the value of b being determined in each case by the observed 500-hour results. The correlation indices for the three groups were 0.975, 0.996 and 0.948, respectively. Curves for all three groups are plotted in Fig. 2. As an additional check on the general fit of the curve $y=99b^x$, several types of 5,000-hour tubes were analyzed for correlation between observed data and the straight line based on the 500-hour percentage for each type. The index of correlation with the straight line was 0.986 for triode oscillators, 0.968 for radio-frequency pentodes and 0.976 for audio and video power amplifiers. Observed and calculated curves for each type are shown in Fig. 3.

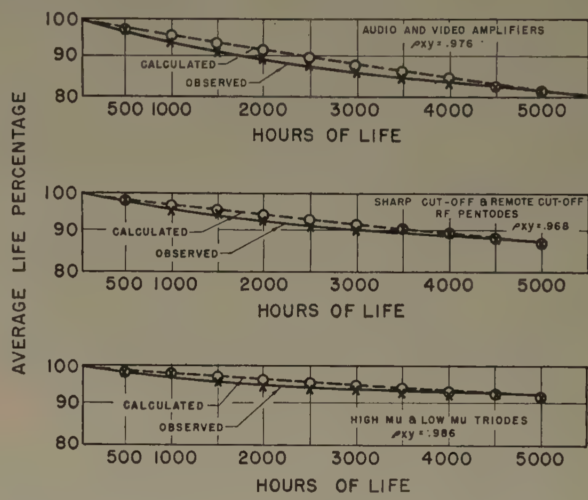

Fig. 3—Life survival curves: 5,000-hour tubes. Curves showing the correlation of observed and calculated percentages for several types of Premium Subminiature Tubes. Life-test conditions for types indicated were as follows:

	Ef	*Eb*	*Rk*	*Ec2*	*Rg*	*Ehk*
audio beam power tube	6.3	110	270Ω	110	500K	117Vrms
video amplifier	6.3	150	100Ω	100	500K	117Vrms
sharp cutoff pentode	6.3	100	150Ω	100	1 meg	117Vrms
semi-remote cutoff pentode	6.3	100	120Ω	100	1 meg	117Vmrs
high mu triode	6.3	150	680Ω	—	1 meg	117Vrms
low mu triode	6.3	100	150Ω	—	1 meg	117Vrms

The correlation between predicted percentages and actual results was in all cases so high that the equation $y=99b^x$ was accepted as the basis for all future statistical controls to be applied to life tests. This pattern of failure was a definite departure from the normal curve expected, and posed additional problems in the development of statistical controls. Earlier experience with incandescent and fluorescent lamps, and with tungsten filament-type vacuum tubes, indicated a certain wear-out point around which regular sigma limits could be plotted, as on a normal Gaussian type curve. A recent

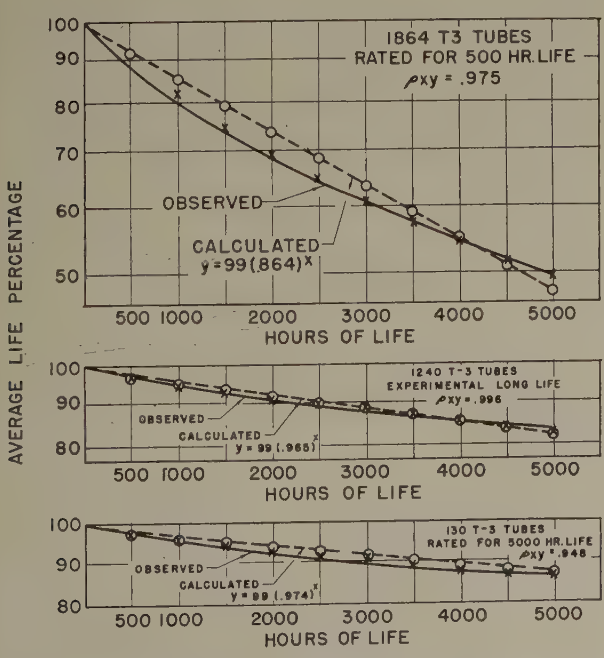

Fig. 2—Life survival curves: $y=99b^x$. Correlation of observed data with curves calculated from 500-hour percentages for three heterogeneous groups of tubes.

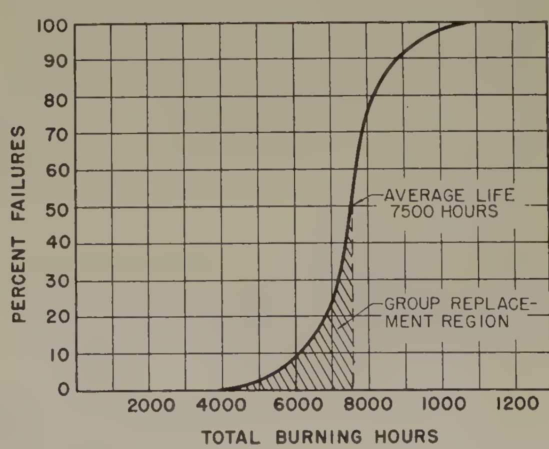

Fig. 4—Life failure curve for fluorescent lamps. Fluorescent lamp life data plotted as a normal Gaussian type curve.

advertising bulletin for improved fluorescent lamps showed this type of curve plotted for failure percentages, with the average life marked at 7,500 hours, and standard deviations of 1,000 hours counted off on either side, as in Fig. 4. With such data, engineers can plan on a minimum life for each lamp, or an optimum replacement point for any group of lamps. Manufacturer's advertisements recommend the replacement of panels of lamps after 5,500 hours, a point at which a maximum of 2.5 per cent of the lamps will have failed. The 20 per cent, or 50 per cent, or 90 per cent failure points could be located just as easily.

Unfortunately, failure data for subminiature vacuum tubes do not follow a normal distribution, and conventional measures of central tendency and dispersion are not applicable to the problem of determining proper control limits. Therefore, it became necessary to devise a system of controls which might be applied to the exponential curve $y=99b^x$.

The first step in the process of setting up controls was to determine, from the 80 per cent—5,000 hour specification and the calculated $y=99b^x$ curve, a minimum limit to be applied at 500 hours. This 500-hour percentage was found to be 96.9 per cent. Then from accumulated data on subminiature tubes, 133 sample life tests of five tubes each were chosen which passed this 96.9 per cent—500-hour limit. Of these tests, not one failed to meet an 80 per cent—5,000-hour limit at the conclusion of the specified life test. These results led naturally to the conclusion that the minimum limit calculated was well chosen, and that the modified 500-hour test would serve as an adequate control on 5,000-hour quality.

Although the choice of a 500-hour limit was the solution to the original problem, it raised a new question of equal importance to manufacturer and customer. This new topic was the probability of release of a lot of tubes which would fail to meet the 5,000-hour life specification. In order to calculate the range of probability of such an occurrence, the five-tube sample life tests mentioned above were used to plot a frequency distribution of 5,000-hour percentages, as shown in Fig. 5. The average 5,000-hour percentage was found to be 89.4 per cent, with 2-sigma limits of 80.3 per cent and 98.5 per cent. These limits on the sample distribution were changed to limits on the universe or parent population by use of the formula

$$\sigma_{\text{sample}} = \left(\frac{N-1}{N}\right)^{1/2} \sigma_{\text{universe}},$$

resulting in new 2-sigma limits of the universe of 80.1 per cent and 98.7 per cent. Statistically speaking, 95 per cent of all released lots of tubes will fall within these limits. Conversely, 2.5 per cent of all released lots may fall on either side of these limits. To use a phrase common to all fields in which quality control is applied, it seems safe to assume an acceptable quality level (AQL) of 2.5 per cent at 5,000 hours.

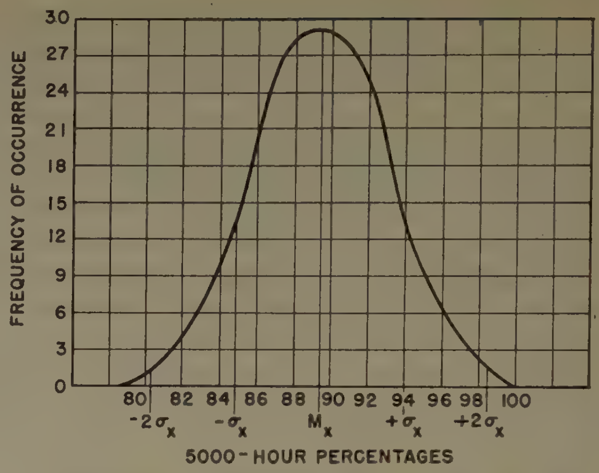

Fig. 5—Distribution of 5,000 hour percentage: 133 sample life tests, 5 tubes each. Range of average life percentages at 5,000 hours for a heterogeneous group of 133 sample life tests.

It would not be reasonable to assume that this life test plan will work equally well for all types of vacuum tubes, made by various manufacturers, until sufficient data to 5,000 hours has been collected and analyzed. The data included in this paper represent only subminiature indirectly heated cathode-type vacuum tubes made at the Kew Gardens Development Laboratory of Sylvania Electric Products Inc. Whether other tube types, or even the same types manufactured elsewhere, would produce equivalent results is a question which only the comparison of actual data will answer. Experience shows that the plan may be applied only to tubes which are designed for long-life operation, are conservatively rated, and are carefully controlled during production. To the writer's knowledge, there has been only one other published indication of an exponential curve of life percentages versus time, a life curve on repeater tubes published by the Bell Telephone Laboratories.[3] It is to be hoped that long-life data may be collected throughout the industry, and that universal life test specifications may be agreed upon by manufacturers and customers. The 500-hour test specification included in this paper was developed with the co-operation of the Bureau of Ships, the chief customer for the tube types represented, and was accepted by them for these particular types as manufactured in Kew Gardens, L. I., N. Y.

There remains one important point not yet mentioned: what kind of guarantee can be given to the customer? What will the manufacturer do if a group of tubes fails to meet the specified life rating in actual operation? Unfortunately, there is no satisfactory answer as yet. For subminiature long-life tubes, there are certain applications, such as hermetically sealed assemblies, where replacement of tubes is impossible. In many other applications for which subminiature tubes have been specially designed, replacement is difficult and expensive. In some cases, the failure of a tube may

[3] D. K. Gannett, "Determination of the average life of vacuum tubes," *Bell Tel. Lab. Rec.;* August, 1940.

TABLE I

STATISTICAL EVALUATION OF LIFE EXPECTANCY OF VACUUM TUBES DESIGNED FOR LONG-LIFE OPERATION

1,864 T-3 Vacuum Tubes Rated for 500 Hours Life

Hours of Life	*Number of Failures*	*Average Life Percentage at End of Period*
0– 500	298	92.0
500–1,000	153	82.0
1,001–1,500	113	74.8
1,501–2,000	84	69.2
2,001–2,500	72	64.8
2,501–3,000	75	61.0
3,001–3,500	43	57.2
3,501–4,000	65	54.8
4,001–4,500	36	51.4
4,501–5,000	53	49.5

cause the destruction of the entire unit. What the customer requires, therefore, is not a replacement guarantee for tubes which prove unsatisfactory, but a certain degree of assurance of reliability of operation. The plan proposed is an illustration of the application of statistical analysis to this difficult quality control problem. Although this plan may not be the perfect answer to the customers' requirements, it is a step in the right direction. It is at least a foundation for future improvements.

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(3) C. Eisenhart, M. W. Hastay, and W. A. Wallis, "Selected Techniques of Statistical Analysis," McGraw-Hill Book Co., Inc., New York, N. Y.; 1947

Magnetic Recording with AC Bias*

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Summary—**The function of alternating-current (ac) bias in magnetic recording is analyzed in a manner similar to that used to explain amplitude modulation. Certain simplifying assumptions are made to facilitate manipulation of mathematical expressions. The analytical results are compared with experimental observations of harmonic distortion, amplitude of fundamental, spurious recorded frequencies, frequency response, difficulty of erasure, and the like.**

INTRODUCTION

IN A MODULATOR for amplitude modulation (AM) radio transmission, an audio frequency and a much higher "carrier" frequency are combined in a nonlinear impedance. The output contains the two original frequencies, their sum, their difference, certain harmonics of each depending upon the character of the nonlinear impedance, and sums and differences of harmonics and fundamentals. A value of carrier amplitude must be selected for the particular nonlinear element to provide linearity and sufficient output in the desired band, which includes the carrier frequency and the carrier-audio sum and difference frequencies. A band-pass filter (tank circuit) is provided to attenuate undesired frequencies. The need for selecting a particular carrier amplitude is most obvious in the case of grid modulation.

In like manner, the action of alternating current (ac) bias in magnetic recording may be explained. The desired audio frequency and a much higher "bias" frequency are simultaneously fed into a nonlinear recording system. The recording contains the audio frequency, the bias frequency, and in addition to these, certain harmonics of each, and sums and differences of harmonics and fundamentals. A value of bias amplitude must be selected for the particular nonlinear recording characteristic to provide linearity and sufficient output in the desired audio band. Self-demagnetization in the recording medium and limited playback head resolution provide a low-pass filter which attenuates undesired (higher than audio) frequencies.

With the shapes of nonlinear recording characteristics in general use, this "bias" technique provides greatly reduced harmonic distortion of the audio, as compared to direct current (dc) bias or no bias.

This technique is capable of improving linearity of response for desired frequencies in a variety of nonlinear systems, whether for transmission or recording.

SCHEME FOR ANALYSIS

The transfer characteristics for a magnetic recording material is the B_r–H curve,[1] (see Fig. 1). Such a curve may be plotted from data taken by single dc exposures or from data taken in the symmetrically cyclically magnetized condition (SCMC). Curves plotted in these two ways are very similar, though not identical. A convenient set of measuring equipment for the SCMC case has been described by Wiegand and Hansen.[2]

* Decimal classification: R365.35. Original manuscript received by the Institute, April 28, 1950; revised manuscript received, August 31, 1950.

† Armour Research Foundation, Illinois Institute of Technology, Chicago, Ill.

[1] Marvin Camras, "Graphical analysis of linear magnetic recording using high-frequency excitation," PROC. I.R.E., vol. 37, pp. 569–573; May, 1949.

[2] D. E. Wiegand and W. W. Hansen, "A 60-cycle hysteresis loop tracer for small samples of low-permeability material," *Trans. AIEE*, vol. 66; 1947.

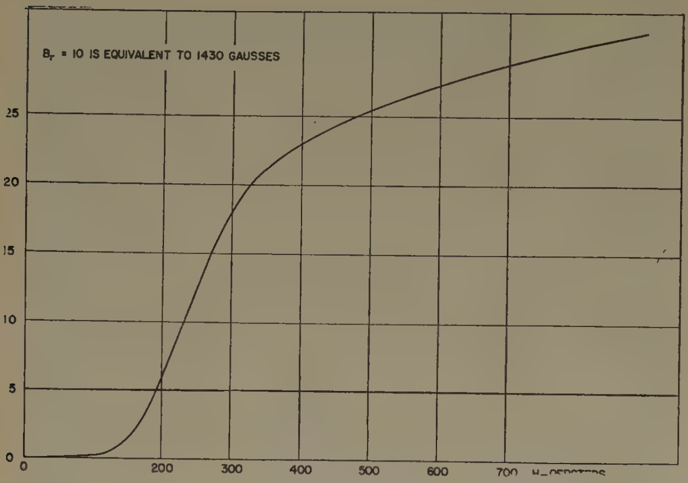


Fig. 1—Retained flux density versus peak magnetizing field. (B_r–H curve) for wire no. 449.

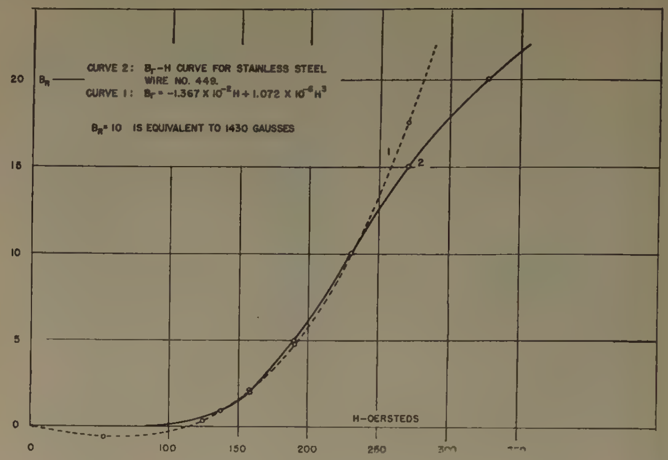


Fig. 2—Cubic or third-order fit to the B_r–H curve.

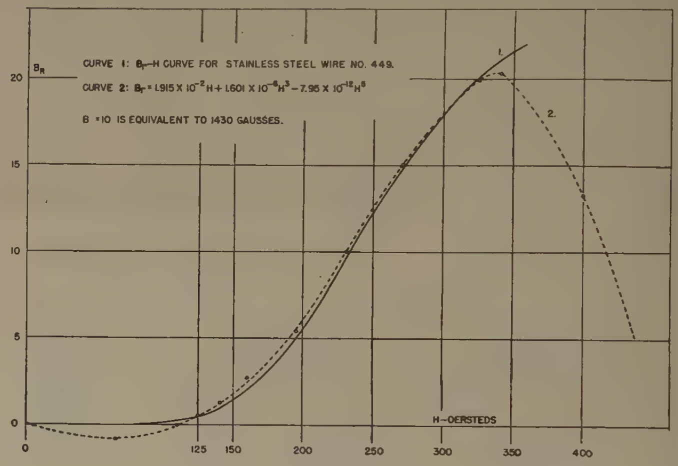


Fig. 3—Fifth-order fit to the B_r–H curve.

In a practical magnetic recorder, each element of length of the recording medium is subjected to substantially a single value of audio field and to several decaying cycles of bias field. The similarity of single exposure and SCMC B_r–H curves permits us to assume that each element of length of the recording medium is subjected to a single instantaneous value of both audio and bias. We shall later discuss phenomena inconsistent with this assumption.

It would be desirable to find an expression $B_r = F_1(H)$ which closely fits the B_r–H curve, assume constant velocity of the recording medium, set in $H = F_2(t)$, and manipulate the resulting equation into forms which can be physically interpreted.

An expression which provides a good fit is

$$B_r = \frac{2B_{rs}}{\pi}(\tan^{-1} K_1H - K_1He^{-K_2H^2}).$$

However, this expression is quite resistant to a desired kind of manipulation, so we resort to a series,

$$B_r = -K_1H + K_2H^3 - K_3H^5. \tag{1}$$

Fits provided by two and three terms are illustrated in Figs. 2 and 3. Use of more terms should improve the fit, but the difficulties of manipulation then become severe. In the following, the general expressions are now and then reduced to numerical values for the particular curves shown in order to insure that conclusions drawn are limited to the regions of reasonably good curve fits.

Case I. Ac Bias

Assuming a sine wave of audio, $A \sin at$, and a sine wave of bias,

$$B \sin bt, \text{ we let } H = A \sin at + B \sin bt \tag{2}$$

and substitute this in (1).

$$\begin{aligned} B_r = &- K_1[A \sin at + B \sin bt] \\ &+ K_2[A \sin at + B \sin bt]^3 \\ &- K_3[A \sin at + B \sin bt]^5. \end{aligned} \tag{3}$$

To simplify notation let $X = A \sin at$

$$Y = B \sin bt,$$

then

$$\begin{aligned} B_r = &- K_1[X + Y] + K_2[X^3 + 3X^2Y + 3XY^2 + Y^3] \\ &- K_3[X^5 + 5X^4Y + 10X^3Y^2 \\ &+ 10X^2Y^3 + 5XY^4 + Y^5]. \end{aligned} \tag{4}$$

The following identities are useful:

$$A^3 \sin^3 at = A^3 \frac{3 \sin at - \sin 3at}{4}, \tag{5}$$

$$B^3 \sin^3 bt = B^3 \frac{3 \sin bt - \sin 3bt}{4}, \tag{6}$$

$$A^5 \sin^5 at = A^5 \frac{\sin 5at - 5 \sin 3at + 10 \sin at}{16}, \tag{7}$$

$$B^5 \sin^5 bt = B^5 \frac{\sin 5bt - 5 \sin 3bt + 10 \sin bt}{16}, \tag{8}$$

$$A^2B \sin^2 at \sin bt$$

$$= A^2B \left\{\frac{\sin bt}{2} - \frac{1}{4}[\sin(2a + b)t - \sin(2a - b)t]\right\}, \tag{9}$$

$AB^2 \sin at \sin^2 bt$

$$= AB^2 \left\{\frac{\sin at}{2} - \frac{1}{4}\left[\sin(2b+a)t - \sin(2b-a)t\right]\right\}, \quad (10)$$

$$A^3B^2 \sin^3 at \sin^2 bt = A^3B^2\{3/8 \sin at - 1/8 \sin 3at - 3/16[\sin(a-2b)t + \sin(a+2b)t] + 1/16[\sin(3a-2b)t + \sin(3a+2b)t]\}, \quad (11)$$

$$A^2B^3 \sin^2 at \sin^3 bt = A^2B^3\{3/8 \sin bt - 1/8 \sin 3bt - 3/16[\sin(b-2a)t + \sin(b+2a)t] + 1/16[\sin(3b-2a)t + \sin(3b+2a)t]\}, \quad (12)$$

$$A^4B \sin^4 at \sin bt = A^4B\{3/8 \sin bt - 1/4[\sin(b-2a)t + \sin(b+2a)t] + 1/16[\sin(b-4a)t + \sin(b+4a)t]\}, \quad (13)$$

$$AB^4 \sin at \sin^4 bt = AB^4\{3/8 \sin at - 1/4[\sin(a-2b)t + \sin(a+2b)t] + 1/16[\sin(a-4b)t + \sin(a+4b)t]\}. \quad (14)$$

Although it is not desirable, a recorder might be designed for an upper audio limit of 15,000 cps ($a = 2\pi \times 15{,}000$) while using a bias frequency of 40,000 cps ($b = 2\pi \times 40{,}000$). Visualizing such values, we may now inspect the above identities to see what terms occur in the audio region. Self-demagnetization and limited playback resolution will attenuate short wavelengths (high frequencies) and permit us to drop such terms, provided we have chosen an appropriate recording medium speed.

We can see that angular frequencies a, $3a$, $5a$, $(b-2a)$, and $(b-4a)$ fall in the audio range. If we had chosen a higher bias frequency, say 80,000 cps, the last two difference frequencies would be outside the audio band and therefore negligible. Practical design experience has also shown that a bias frequency at least 5 times the highest audio frequency admitted to the recording head is desirable to avoid distortion from audible beat frequencies. Slightly lower bias frequency may be satisfactory if the recording current amplitude is very small at the upper audio frequencies, as is often true in sound recording.

Let us now decide to use a bias frequency sufficiently high to make these beat notes negligible. Then our expansion of B_r contains terms in a, $3a$, and $5a$.

$$B_r = \left[-AK_1 + \frac{3A^3K_2}{4} + \frac{3AB^2K_2}{2} - \frac{5A^5K_3}{8} - \frac{15A^3B^2K_3}{4} - \frac{15AB^4K_3}{8}\right] \sin at + \left[\frac{5A^5K_3}{16} + \frac{5A^3B^2K_3}{4} - \frac{A^3K_2}{4}\right] \sin 3at - \frac{A^5K_3}{16} \sin 5at. \quad (15)$$

Let us call the coefficient of sin at, C_a, the coefficient of sin $3\,at$, C_{3a}. For zero third harmonic,

$$C_{3a} = 0$$

$$\frac{5A^5K_3}{16} + \frac{5A^3B^2K_3}{4} - \frac{A^3K_2}{4} = 0$$

$$20K_3B^2 = 4K_2 - 5A^2K_3$$

$$B^2 = \frac{K_2}{5K_3} - \frac{A^2}{4}, \quad (16)$$

which indicates that the amount of bias for zero third harmonic varies with audio amplitude, decreasing as the audio increases.

$$\text{As } A \to 0, \ B^2 \to \frac{K_2}{5K_3} \text{ or } 0.2\frac{K_2}{K_3}. \quad (17)$$

$B = 200$ oersteds for the material shown in Fig. 3, which is in the region of a reasonable curve fit.

It is also of interest to note that the sign of C_{3a} reverses as B increases from near zero to above the value for $C_{3a} = 0$.

Let us now investigate C_a, seeking a maximum value.

$$C_a = -AK_1 + \frac{3A^3K_2}{4} + \frac{3}{2}AB^2K_2 - \frac{5A^5K_3}{8} - \frac{15}{4}A^3B^2K_3 - \frac{15}{8}AB^4K_3 \quad (18)$$

$$\frac{dC_a}{dB} = 3ABK_2 - \frac{15}{2}A^3BK_3 - \frac{15}{2}AB^3K_3$$

when $d(C_a)/dB = 0$ (disregarding root at $B = 0$)

$$15B^2K_3 = 6K_2 - 15A^2K_3$$

$$B^2 = 0.4\frac{K_2}{K_3} - A^2, \quad (19)$$

which indicates that the amount of bias for maximum C_a decreases when A increases. When $A \to 0$,

$$B^2 \to 0.4\frac{K_2}{K_3}$$

$$B = \sqrt{0.4\frac{K_2}{K_3}} \quad (20$$

which is $\sqrt{2}$ the value of B for $C_{3a} = 0$.

$B \doteqdot 284$ oersteds for the material of Fig. 3.

If we had chosen to drop the fifth order term from (1), we would have

$$B_r = -K_1H + K_2H^3$$

$$\frac{dB_r}{dH} = -K_1 + 3K_2H^2$$

$$\frac{d^2B_r}{dH^2} = 6K_2H.$$

The only inflection point is at $H = 0$, and the saturation phenomenon is not represented. Applying the sub-

stitution $H=A \sin at+B \sin bt$, we get an expression for B_r involving fewer angular frequencies than in the fifth-order case. Assuming a high bias frequency and low pass filtering, we get

$$B_r = \left[-AK_1 + \frac{3}{4}A^3K_2 + \frac{3}{2}AB^2K_2\right]\sin at - \frac{A^3K_2}{4}\sin 3at. \tag{21}$$

If we seek a value for zero third harmonic, we find it only at $A=0$, which is a trivial result. When we seek the value of B for maximum fundamental, we find it at $B=\pm\infty$, which is due to failure to include the saturation phenomenon.

It is, therefore, apparent that at least 3 terms of (1) must be taken into account to get physically significant results.

Case II. No Bias Compared to Ac Bias

If the signal is only audio, $H=A \sin at$, we have from (1)

$$B_r = -K_1A\sin at + K_2A^3\sin^3 at - K_3A^5\sin^5 at \tag{22}$$

or

$$B_r = -K_1A\sin at + K_2A^3\frac{3\sin at - \sin 3at}{4} - K_3A^5\frac{\sin 5at - 5\sin 3at + \sin at}{16}$$

$$B_r = \left[-K_1A + \frac{3}{4}K_2A^3 - \frac{5}{8}K_3A^5\right]\sin at + \left[-\frac{K_2A^3}{4} + \frac{5}{16}K_3A^5\right]\sin 3at - K_3A^5\sin 5at. \tag{23}$$

Comparing this with (15), we find all terms common except those involving B in (15). Using ac bias as against no bias, C_a is increased in the case of ac bias by the addition of the quantity

$$\left(\frac{3}{2}AB^2K_2 - \frac{15}{4}A^3B^2K_3 - \frac{15}{8}AB^4K_3\right)$$

and C_{3a} is increased by the addition of $(5/4)A_3B^2K_3$. In order for any benefit to result from the use of ac bias, an improvement in linearity must result, that is, Ca must be more nearly proportional to A. This will occur if the value of terms in C_a which involve A^3 or A^5 are reduced by the use of bias.

Let us test

$$\frac{3}{4}K_2A^3 = \frac{15}{4}B^2K_3A^3 \tag{24}$$

$$K_2 = 5K_3B^2$$

$$B^2 = 0.2\frac{K_2}{K_3}.$$

This value of B will eliminate the A^3 term from C_a, and it is the value of bias for zero third harmonic when $A\rightarrow 0$, deduced in (17). Assuming this value of bias, we have

$$C_a = -AK_1 + \frac{3}{4}A^3K_2 + \frac{3}{10}\frac{AK_2^2}{K_3} - \frac{5}{8}A^5K_3 - \frac{3}{4}A^3K_2 - \frac{3}{40}\frac{AK_2^2}{K_3},$$

$$C_a = -AK_1 + \frac{9}{40}\frac{AK_2^2}{K_3} - \frac{5}{8}A^5K_3. \tag{25}$$

It appears that good linearity and high sensitivity with ac bias will result if K_3 is very small compared to K_2, and if K_1 is small compared to K_2^2/K_3. However, as K_3 decreases, the bias required increases.

It seems probable that a seventh-order equation would provide a term involving BA^5 which would reduce the A^5 term in C_a.

Case III. Dc Bias on a Neutral Medium

If the signal is audio plus a dc bias, $H=D+A \sin at$, and

$$B_r = -K_1[D + A\sin at] + K_2[D + A\sin at]^3 - K_3[D + A\sin at]^5, \tag{26}$$

$$B_r = -K_1D - K_1A\sin at + K_2[D^3 + 3AD^2\sin at + 3A^2D\sin^2 at + A^3\sin^3 at] - K_3[D^5 + 5AD^4\sin at + 10A^2D^3\sin^3 at + 10D^2A^3\sin^3 at + 5DA^4\sin^4 at + A^5\sin^5 at]. \tag{27}$$

The following additional identities are now useful:

$$\sin^2 at = \frac{1 - \cos 2at}{2}, \tag{28}$$

$$\sin^4 at = \frac{1}{8}(3 - 4\cos 2at + \cos 4at). \tag{29}$$

Substituting in (26),

$$B_r = -K_1D + K_2D^3 + \frac{3}{2}K_2DA^2 - K_3D^5 - 5K_3D^3A^2 - \frac{15}{8}K_3DA^4 + \left[-K_1A + 3K_2D^2A + \frac{3}{4}A^3K_2 - 5K_3D^4A - \frac{15}{2}K_3D^2A^3 - \frac{5}{8}K_3A^5\right]\sin at + \left[-\frac{3}{2}K_2DA^2 + 5K_3D^3A^2 + \frac{5}{2}K_3DA^4\right]\cos 2at + \left[-\frac{A^3K_2}{4} + \frac{5}{2}K_3D^2A^3 + \frac{5}{16}K_3A^5\right]\sin 3at$$

$$+\left[-\frac{5}{8}K_3DA^4\right]\cos 4at$$

$$-\frac{K_3A^5}{16}\sin 5at. \qquad (30)$$

The condition for $C_{2a}=0$ is

$$\frac{3}{2}K_2 = 5K_3D^2 + \frac{5}{2}K_3A^2, \qquad (31)$$

$$D^2 = \frac{3K_2 - 5K_3A^2}{10K_3}, \qquad (32)$$

which indicates that the value of dc bias for zero second harmonic varies with audio amplitude.

$$\text{As} \quad A \to 0,\ D^2 \to \frac{3K_2}{10K_3}. \qquad (33)$$

This is the inflection point of (1), or $D=245$ oersteds, for the material shown in Fig. 3.

The condition for $C_{3a}=0$ is

$$\frac{5}{2}K_3D^2 + \frac{5}{16}K_3A^2 = \frac{K_2}{4}, \qquad (34)$$

$$D^2 = \frac{4K_2 - 5K_3A^2}{40K_2} \quad (D \text{ for zero third varies with } A) \qquad (35)$$

as $A\to 0$ $D^2\to(K_2/10K_2)$ so that D differs by a factor of $\sqrt{3}$ from the value of D for $C_{2a}=0$, indicating that no one value of dc bias minimizes even and odd harmonics. However, no spurious beat frequencies are encountered as in ac bias.

This analysis does not apply directly to the use of dc bias on previously saturated recording medium. Equation (1) does not apply to this situation.

Phenomena Consistent with the Analysis

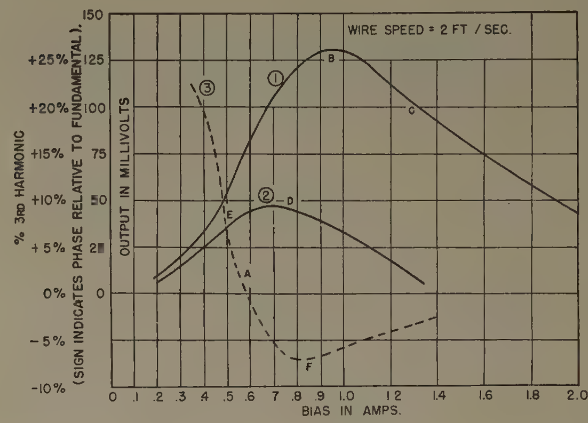
Fig. 4 shows some experimental data on amplitude and phase of third harmonic and fundamental as a function of bias amplitude, using a small constant value of audio current. The recording wire used in these experiments is the same as was used for the B_R–H curve shown in Fig. 1. The following points of agreement between the data of Fig. 4 and the analysis are found:

1. The change in sign of the third harmonic as ac bias is increased, and the existence of a value of ac bias which provides zero third harmonic.
2. The existence of a maximum value for the coefficient of the fundamental at an ac bias amplitude higher than that required for zero third harmonic.

In other experimental work, the following additional points of agreement with the analysis have been found:

3. The existence of beat notes between the ac bias frequency or its harmonics and harmonics of high audio frequencies. The frequencies $(b-2a)$, $(b-4a)$, and $(2b-5a)$ have been identified. The last of these would appear in the analysis if a seventh-order series were used for (1).

PLAYBACK WAVE FORMS (SOLID) AND HARMONIC ANALYSIS (DOTTED). THESE PATTERNS INCLUDE THE DIFFERENTIATING EFFECT OF THE PLAYBACK PROCESS.



FUNDAMENTAL AND 3RD. HARMONIC VS. BIAS

Fig. 4—Experimental results.
① 100 cycles: fundamental output versus bias.
② 5,000 cycles: fundamental output versus bias.
③ 100 cycles: per cent third harmonic versus bias.
Audio recording current constant for all data.

4. The change with audio amplitude of the value of ac bias required for maximum amplitude of fundamental.
5. The change with audio amplitude of the value of ac bias required for zero third harmonic.
6. The occurrence of even-order harmonics in the presence of dc bias, dc bias may be a result of direct current in the recording head, or residual magnetization in the recording head or residual magnetization in the medium as it reaches the recording head.
7. The improvement in linearity with appropriate amplitude of ac bias.

Some Phenomena Inconsistent with the Assumptions or Outside the Scope of the Analysis

1. *Frequency Response*

The analysis offered does not show that relative high-frequency response falls off more rapidly in the case of ac bias than with dc bias on neutral medium. This does occur, however, and it is ascribed to an erasing or aging action in the case of ac bias. An element of length of the medium is exposed to several decaying ac cycles. This is more effective in reducing the amplitude of short wavelengths than in reducing the amplitude of long wavelengths. Short magnets are more easily demagnetized than long magnets. This aging action is desirable in that the recording is more stable if ac bias is used.

2. *Difficulty of Erase*

Nothing in the analysis ascribes a different quality to a B_r achieved with small audio and large ac bias than to

the same value of B_r achieved with larger audio and less ac bias. Herr, Murphy, and Wetzel[3] have shown that the signal resulting from small audio and large bias is the more difficult to erase. In the charging of permanent magnets for meters and transducers, it is found that more stable (more difficult to demagnetize) magnets result from initially saturating the magnet with dc, and then "aging" it down to the desired B_r with ac fields, than by only exposing it to sufficient dc to bring it up to the desired B_r from the demagnetized condition. This is analogous to the results described by Herr, Murphy, and Wetzel.

3. *Even Harmonics Resulting from Asymmetry of Bias Wave Form*

The analysis does not account for even harmonics resulting from asymmetry of the bias wave form.

Textbooks on magnetic phenomena describe the fact that the residual induction in a piece of iron which has been exposed to a varying unidirectional field is determined by the peak value of the field, independent of the time variations of the field.

Thus, so far as the recording process is concerned, asymmetrical bias wave form has a dc effect, even when there is no dc component in the bias field itself. The field has zero dc component if the area bounded by the curve on each side of zero is equal. The recorded effect is quite independent of this area, and sensitive to peak values. Quantitative evaluation of this phenomenon is complicated by the fact that ac bias swings past zero, into the opposite polarity.

4. *Variation of Bias with Wavelength*

From the analysis presented it would appear that for small audio amplitudes, maximum fundamental amplitude and zero third harmonic appear at particular values of bias, independent of audio frequency or wavelength. This is not true, actually, for at very short wavelengths no appreciable third harmonic is apparent at any level of bias, because of low pass filtering effects (self-demagnetization, playback gap effect, eddy current effects, and electrical circuit filtering effects). At frequencies well below the frequency of maximum response, the third harmonic is accentuated in playback due to rising frequency response. Playback voltage is proportional to the rate of change of B_r, rather than proportional to B_r itself, thus producing rising frequency response in the range of wavelengths which suffer little, or constant, self-demagnetization.

Erasing or aging effectiveness of the bias is also greater at short wavelengths.

[3] R. Herr, B. F. Murphy, and W. W. Wetzel, "Some distinctive properties of magnetic-recording media," *Jour. Soc. Mot. Pic. Eng.*, vol. 52; January, 1949.

5. *Decrease of Distortion at Large Bias Amplitudes*

The analysis presented does not account for the decrease in distortion at bias amplitudes larger than that at point *F* in Fig. 4. High bias values are outside the region of good fit of even the fifth-order series. It is believed that, if many more terms were used, this decrease in third harmonic would appear.

Optimum Bias Frequency

For thin recording wires and tapes, it is desirable that the bias frequency be as high as practical considerations will permit. In designing a recorder, one usually wishes to use the same oscillator for erase and bias. Eddy current losses in the erase head cause the designer to consider the use of low erase-bias frequencies. The designer usually decides upon an erase-bias frequency slightly lower than five times the upper audio limit of his system as the best compromise.

Optimum AC Bias Amplitude

It has been shown that the bias amplitudes for maximum fundamental output and zero third harmonic are not the same, and further that the bias amplitudes for maximum fundamental and zero third harmonic vary with audio amplitude. Also, the bias amplitude for any of these conditions varies with the magnetic materials and geometries of heads, tapes, and wires. Thus it is not surprising that there should be a variety of schemes favored by various people for setting the bias amplitude.

For low-cost, low tape-speed equipment, designers often set the bias amplitude for maximum sensitivity at short wavelengths, tolerating the accompanying distortion of long waves on the tape. In such equipment the harmonic distortion in the recording process is often negligible compared to the distortions present in the amplifier.

In high-cost, high-quality equipment, a larger bias amplitude is usually favored. This is much greater than the bias for zero third harmonic at medium and high audio frequencies, and is in the region of low distortion to the right of point *F* in Fig. 4. Operation in this region provides low distortion at all wavelengths and some reduction of relative playback amplitude at short wavelengths. The latter is corrected by the use of a sufficiently high tape speed and suitable equalizer circuits.

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Representations of Speech Sounds and Some of Their Statistical Properties*

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Summary—**Spectrographic analysis, autocorrelation, and infinite clipping are considered as methods of transforming and analyzing speech sounds with the object of obtaining simple representations without excessive loss of intelligibility. Although not mathematically equivalent in the exact sense, these methods when properly approximated are shown to provide parameters of statistical nature that are simply related. It is conjectured that the parameters describe some essential elements of speech sounds and are statistically invariant. Some experimental results are included.**

THERE HAS ALWAYS BEEN a need of pictorial or other physical representation ever since man began to analyze speech sounds. The objective may be twofold. One is to find an accurate representation so that every detailed characteristic of the original sound, including naturalness and emotional content, is retained. The other is to obtain as simple a representation as possible without excessive loss of intelligibility. This paper is concerned with the second type.

Speech compression and sensory replacement are two examples of the fields in which application can be made of the representations to be considered. There is considerable interest in both of these fields at the present time as evidenced by current literature and development work by the Bell Laboratories and others in the vocoder field;[1-4] and RCA,[5] Haskins Laboratories,[6] and others[7,8] in the sensory-replacement field.

A problem common to both fields is that of identifying the essential elements for intelligibility as distinguished from the redundant information contained in the original sound. It is conjectured that these essential elements are statistically invariant because of the structural similarity of vocal mechanisms and the tendency of speakers to conform through the process of learning. Names that have been suggested for these essential elements are "gesture," "articulation" and "modulation."[9,10] The problem is to identify and extract them from the original speech sound.

It is the purpose of this paper to compare some methods of analyses and representations which are prevalent, and to attempt to derive therefrom some statistical parameters which may retain part of the invariant essential elements. One method would be to obtain the representation directly from the vocal or hearing mechanisms; for example, to take pictures showing the movements of the articulators[11,12] (vocal chord, lips, tongue, and the like) or pictures, if possible, showing the responses of the tremendous number of auditory nerves.[13] The anatomical and neurological techniques involved are, however, not quite within the reach of a communication engineer. This paper, therefore, will be confined to the acoustic aspects of speech sounds.

A commonly used visual representation of speech sounds is the oscillogram, or the intensity-time graph. The intensity may correspond to the pressure of the original sound or the voltage derived from a microphone. Experience with oscillograms indicates that while simple in appearance, they are very complicated in analysis. Therefore, before invariant elements can be obtained, it is necessary to transform the intensity-time function of the sound to a more suitable form. Since many sounds are periodic, it is natural to introduce the concept of frequency and the use of Fourier analysis, the representation in this case being the spectrogram.[14] Some sounds, however, like fricatives, do not possess a high degree of periodicity; yet, they are not entirely random noise. Spectrographic representations of these sounds are not very satisfactory. Since speech sounds, including fricatives, possess a certain degree of continuity, the correlation between intensities of successive intervals is not insignificant. This type of

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correlation can be expressed by a plot of the autocorrelation function $\phi(\tau)$ versus τ.[15,16] For purely random noise, $\phi(\tau)$ drops instantaneously as τ exceeds zero. For periodic functions, $\phi(\tau)$ varies with the same periodicity. For fricatives with hidden periodicity, $\phi(\tau)$ may fall rather rapidly, followed by small ripples. The spectrogram and the autocorrelationgram are two transforms of the oscillogram and have been considered previously by others.

A representation which heretofore has not received much consideration is clipped speech.[17] It is reported that high intelligibility is retained by replacing the original speech wave with rectangular waves having the same zero crossings. There are, of course, other types of representations, such as pulse-code sampling,[18] complex-frequency mapping,[19] and so forth; however, this paper is concerned with only the four types mentioned above because of the apparent feasibility of experimental attack.

Fig. 1 shows these four representations: the oscillogram, the spectrogram, the autocorrelation function, and clipped speech. Mathematically, if we denote the original sound by $f(t)$, the spectrogram is the Fourier transform of $f(t)$, denoted by $F(f)$. It contains information concerning both amplitude and phase. Under usual conditions, this transform is reversible as indicated by an arrow pointed both ways. This transform can be used in the direct sense ($\rightarrow$) for analysis, or in the inverse sense ($\leftarrow$) for synthesis. The autocorrelation function $\phi(\tau)$ as shown is defined as the average of the product of the original function $f(t)$ and the function shifted in time $f(t-\tau)$. This transformation is expressed mathematically as

$$\phi(\tau) = \lim_{T\to\infty} \frac{1}{2T} \int_{-T}^{+T} f(t)f(t-\tau)dt. \qquad (1)$$

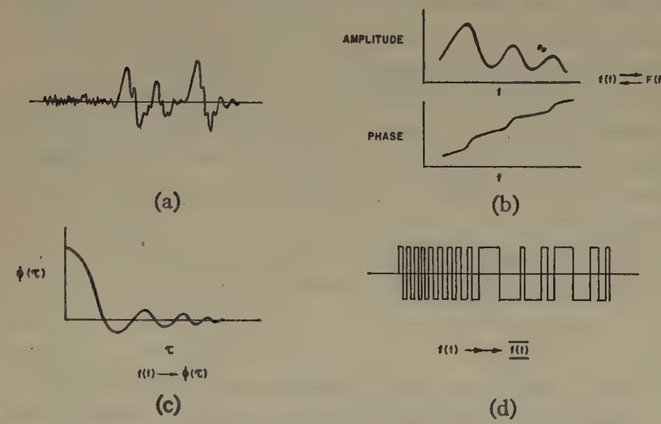

Fig. 1—Some transformations. (a) Original wave form $f(t)$.
(b) Spectrogram

$$F(f) = \int_{-\infty}^{\infty} f(t)e^{-j\omega t}dt.$$

(c) Autocorrelation function

$$\phi(\tau) = \lim_{T\to\infty} \frac{1}{2T} \int_{-T}^{T} f(t)f(t-\tau)dt.$$

(d) Clipped speech $\overline{f(t)}$.

This function, having the dimension of power, can be proved[15] to be the Fourier transform of the power spectrum. Information about phase is not contained in $\phi(\tau)$ and, therefore, the transform is not reversible as indicated by the unidirectional arrow. In the case of clipped speech the transformation is inherently nonlinear, and no mathematical formulation is ventured here, except to assign the symbol $\overline{f(t)}$ to the process. The two arrows pointed the same way indicate the very obvious unidirectional properties of the clipping transformation.

These four representations, although not mathematically equivalent in the exact sense, provide equivalent parameters after proper approximations have been made. These same approximations are necessary for the practical performance of the transformations. Fig. 2 shows the effects of these approximations. The part of the Fourier transform which represents phase is omitted. The magnitude has been changed into power spectral density and lumped into rectangular blocks corresponding to finite bands of frequencies. The spectrogram is considered to be changing with respect to time; hence the inclusion of the time axis. For the autocorrelationgram, the continuous curve is replaced by a finite number of points, corresponding to discrete autocorrelation intervals τ. A time axis t is also introduced.

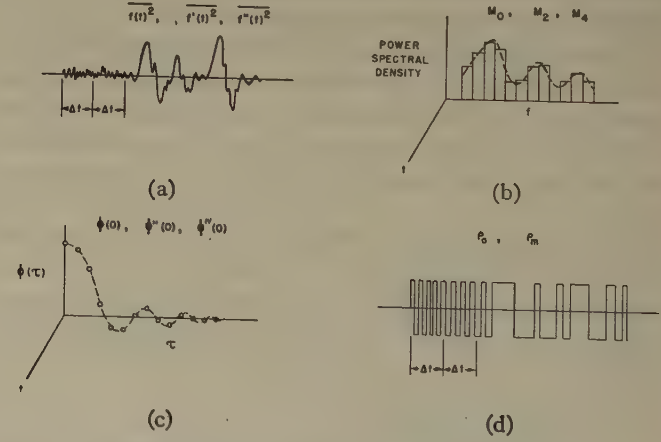

Fig. 2—Approximations in transformations. (a) $f(t)$. (b) $F(f)$.
(c) $\phi(\tau)$. (d) $\overline{f(t)}$.

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[19] W. H. Huggins, "Conjectures Concerning the Analysis and Synthesis of Speech in Terms of Natural Frequencies," an unpublished memorandum, Air Force Cambridge Research Laboratories, Cambridge, Mass., February 9, 1949.

In the case of clipped speech, the approximation has already been considered. It consists of replacing the different intensity levels by only two fixed levels.

In summary, the approximations involved in modifying the three transformations are: (1) the lumping of frequencies in the case of the spectrogram; (2) the use of discrete intervals for autocorrelation time τ; and (3) the use of only two fixed amplitude levels for intensity in the case of clipped speech. Although these approximations involve three different domains—frequency, time, and intensity—it is felt that if they are carried far enough, the results would probably retain about the same amount of information. Licklider and Pollack[17] have shown experimentally that clipped speech retains a high degree of intelligibility. For this reason, and for simplicity, it is appropriate to begin with clipped speech in showing the equivalence of the three transformations.

If the spacing between zeroes is fairly constant in a short interval of time Δt (e.g., of the order of 1/100 second) further simplification is possible. This involves defining a new function ρ_0 giving the average number of zero crossings in the interval Δt (or the zeroes per second). Licklider and Pollack have also reported that a still higher degree of intelligibility is retained if the original speech is differentiated before clipping. In this case, the points of maxima and minima of the original function are retained rather than the points of zero crossing; and a similar density function ρ_m can be defined to give the number of maxima and minima per second. This could be carried still further by defining a third function ρ_i as the density of points of inflection, and so forth. The symbols ρ_0 and ρ_m thus indicate in Fig. 2 two indices which retain certain essential elements and represent a step toward simplification of the original speech sound.

In the study of noise, Rice[20] of the Bell Telephone Laboratories has found that a simple relation exists between the expected density of zeroes $\overline{\rho_0}$ and the averages of the original function and its first derivative. The derivation of this relation as outlined in the Appendix leads to

$$\overline{\rho_0} = k_0 \sqrt{\frac{\overline{f'(t)^2}}{\overline{f(t)^2}}}. \qquad (2)$$

A similar relation can be derived to show that, subject to the same restrictions,

$$\overline{\rho_m} = k_m \sqrt{\frac{\overline{f''(t)^2}}{\overline{f'(t)^2}}}. \qquad (3)$$

In these equations, the proportionality constants k_0 and k_m depend upon the statistical characteristics of the function $f(t)$. For a stationary time series, the ensemble and time averages are indistinguishable. Hence, the mean values $\overline{f(t)^2}$, $\overline{f'(t)^2}$, and $\overline{f''(t)^2}$ are included in Fig. 2 to indicate additional indices leading to simplification of speech sounds.

It can be shown[21] that these averages values are equivalent to the values of the autocorrelation function $\phi(\tau)$ and certain of its first few derivatives with respect to τ at $\tau=0$; thus the symbols $\phi(0)$, $\phi''(0)$, $\phi^{IV}(0)$ appear beside the autocorrelation graph. The relationships between these quantities are given by

$$k_0 \sqrt{\frac{\overline{f'(t)^2}}{\overline{f(t)^2}}} = k_0 \sqrt{\frac{-\phi''(0)}{\phi(0)}} = \bar{\rho}_0 \qquad (4)$$

and

$$k_m \sqrt{\frac{\overline{f''(t)^2}}{\overline{f'(t)^2}}} = k_m \sqrt{\frac{\phi^{IV}(0)}{-\phi''(0)}} = \overline{\rho_m}. \qquad (5)$$

It will be noted that, for smooth functions, the first and the third derivatives at $\tau=0$ are both zero. With the knowledge of these five initial values, the initial portion of the $\rho(\tau)$ curve is fairly well defined.

It can also be shown[15,20] that these initial characteristics of the autocorrelation function are related to the moments of the power spectrum. For example, the moment of zeroth order M_0 is the average power and equals $\phi(0)$. The other two moments M_2 and M_4, are related to the derivatives $\phi''(0)$ and $\phi^{IV}(0)$ as shown by the following equations:

$$k_0 \sqrt{\frac{-\phi''(0)}{\phi(0)}} = 2\pi k_0 \sqrt{\frac{M_2}{M_0}} = \bar{\rho}_0 \qquad (6)$$

$$k_m \sqrt{\frac{\phi^4(0)}{-\phi''(0)}} = 2\pi k_m \sqrt{\frac{M_4}{M_2}} = \bar{\rho}_m. \qquad (7)$$

The symbols M_0, M_2, and M_4 are therefore included along with the spectrogram in the figure.

For the special cases where $f(t)$ is a sine wave and where it is a random function having a Gaussian distribution, it can be proved that k_0 and k_m are equal to $1/\pi$. Application of the relations given by (2) through (7) to a speech sound assumes that it can be regarded as a stationary time series to which a definite distribution can be assigned. Since an exact mathematical formulation of the distribution is very difficult, the extent that this requirement is met can only be conjectured at the present time. However, assuming that these relations do apply to speech sounds, it is suggested that the proportionality constants may not be very much different from $1/\pi$.

The previous discussion has presented three types of analysis from a theoretical standpoint. Certain experimental results will now be described that support the conclusions reached using this approach. Fig. 3 shows ρ_0- and ρ_m-grams, together with the spectrogram,

[20] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. Jour.*, vol. 23, pp. 282–332, July, 1944; and vol. 24, pp. 46–156, January, 1945.

[21] R. Cohen, "Some analytical and practical aspects of Wiener's theory of prediction," Technical Report No. 69, Research Laboratory of Electronics, MIT, Cambridge, Mass., June, 1948.

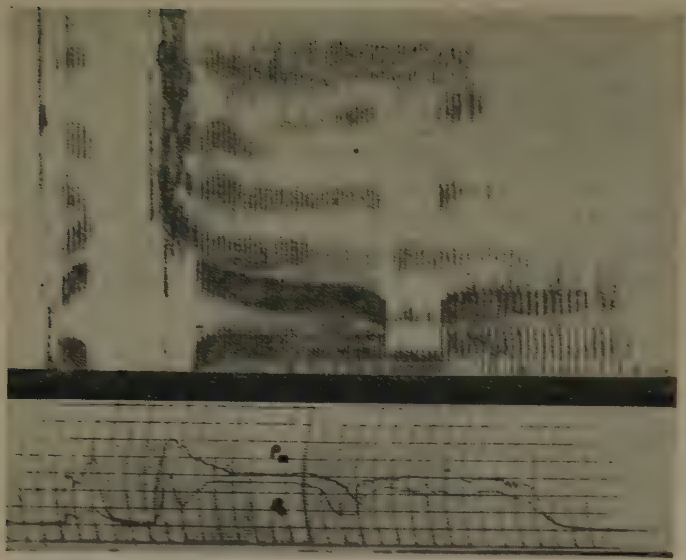


Fig. 3—Spectrogram and P_m and P_0 graphs of "pajama."

for the word "pajama." It is to be noted that there is close similarity between the shapes of the ρ_0- and ρ_m-grams and the first two bars of the spectrogram. The first and second bars of the spectrogram represent the first and second formants, or resonance regions, of the sound. Since the frequency components in the first bar are usually strong enough to cause zero crossings, the ρ_0-gram is a close approximation of this bar. The frequency components in the second resonance region may not be strong enough to cause extra zero crossings, but they will affect the slope of the wave and may, therefore, contribute extra maxima and minima which are included in ρ_m. Fig. 4 shows the graphs of the sounds "i," "ϵ," "α," and "μ" which give additional evidence of this resemblance. Usually vowel and vowel-like sounds such as these, having bars which are clearly distinguishable, will show closer resemblance than other types.

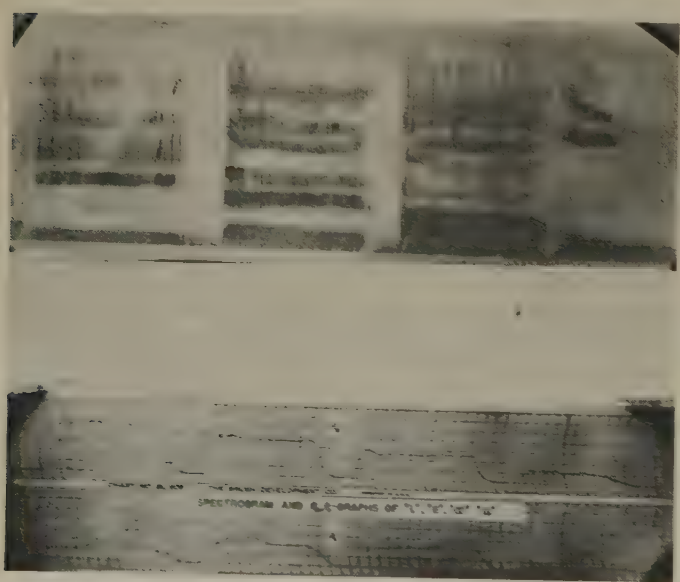


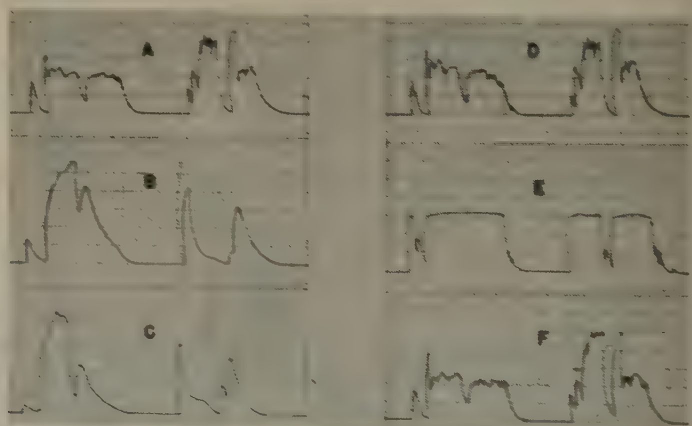
Fig. 4

Fig. 5—Oscillogram of the word "pajama."

A comparison of the ρ_0- and ρ_m-grams of the word "pajama" with the oscillogram of the same word, as shown in Fig. 5, justifies the contention that the former is a considerably simpler representation. Visual inspection of the zero crossings of the oscillogram will show that the general shape of the given ρ_0-gram is correct. The same is true in the case of the maxima and minima.

Some work has been done on the experimental verification of the theoretical relations involving ρ_0, ρ_m, and the other three sets of statistical parameters as applied to speech wave forms. The relationship between ρ_0 and $\overline{f(t)^2}$ and $\overline{f'(t)^2}$ has been checked for some sustained vowel and sustained fricative sounds. This involves the computation of the ratio of the root-mean-square value of $f'(t)$ to that of $f(t)$, i.e., the value of the former normalized with respect to the latter. It is preferable that the mean-square value of the original $f(t)$, that is, $\overline{f(t)^2}$, be kept relatively constant so that $\overline{f'(t)^2}$ will be automatically normalized. This is an inherent property of the types of sounds that have been checked.

For transitional sounds, such as those contained in the words "pajama" and "question," experimental verification is at present impossible since $\sqrt{\overline{f'(t)^2}}$ is dependent upon both amplitude and frequency varia-



(a) (b)

Fig. 6—Graphs for the words "pajama" and "question." (a) Ordinary speech: (A) ρ_0; (B) $\overline{f(t)^2}$; (C) $\overline{f'(t)^2}$. (b) Clipped speech: (D) ρ_0; (E) $\overline{f(t)^2}$; (F) $\overline{f'(t)^2}$.

tions, with the former predominating. Since the $\sqrt{\overline{f(t)^2}}$ itself is dependent only upon the same amplitude variations, the ratio as determined by computation using independent measurements does not yield significant results. It would be necessary to employ some automatic-amplitude-control method to either minimize the changes in amplitude, or to obtain the ratio of the two quantities directly. This effect is illustrated for ordinary speech by graphs (A), (B), and (C) of Fig. 6. Visual examination of (B) and (C) indicates that the amplitude effect predominates. Since clipped speech is an extreme case of amplitude control, data from graphs (D), (E) and (F) should give a better check of the relation in question. Visual examination will show that this is true.

Experimental verification of the other theoretical relations has not as yet been attempted because the autocorrelator and moment computer, while under construction, have not been completed.

Conclusions

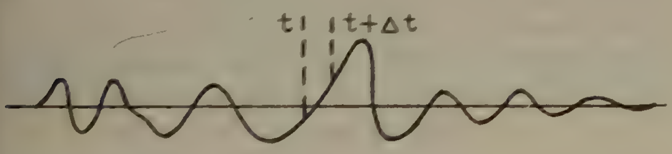
It has been shown that, from a mathematical standpoint, the three methods of analysis considered, namely, spectrographic analysis, autocorrelation, and infinite clipping, provide parameters which are equivalent for certain time functions. Since speech sounds are too complex for mathematical treatment of these types, an experimental attempt at verification of the equivalence for speech sounds is justified. Furthermore, an experimental attempt to determine the degree of invariance of these parameters is advisable. It is hoped that a subsequent paper will report progress currently being made along these lines.

Appendix

I. Mathematical derivations[22]

The Expected Density of Zeroes

Consider a random curve $f(t)$ for which the following figure is a sample of the ensemble. The expected density of



zeroes is given by

$$\overline{\rho_0} = \int_{-\infty}^{\infty} |\eta|\, p(0, \eta) d\eta. \tag{8}$$

In this expression $p(0, \eta)$ is $p(\xi, \eta)$ for $\xi = 0$, where $p(\xi, \eta)$ is the probability density function for the two variables

$$\xi = f \qquad \text{(at time } t\text{)}$$

$$\eta = \frac{df}{dt} = f' \qquad \text{(at time } t\text{);}$$

[22] This derivation is essentially similar to that of Rice. See footnote reference 20.

i.e., the probability that at time t the value of f in the ensemble lies between ξ and $\xi + d\xi$, and the value of f' lies between η and $\eta - d\eta$ is $p(\xi, \eta) d\xi d\eta$. Equation (8) may be proved as follows:

Referring to the figure, in the interval between t and $t - \Delta t$ for all the ensemble, if the slope is positive, i.e.,

$$0 < \eta < \infty, \tag{9}$$

then, in order for the curve to pass through zero in the interval Δt, the following inequality must be satisfied:

$$0 > \; > \eta \Delta t. \tag{10}$$

This is evident from the geometry. The amplitude of ξ must be negative so that the curve will cross the axis with a positive slope η. Its absolute magnitude must be sufficiently small so that the crossing occurs within the interval. The portion of the curve in the interval Δt is assumed to be linear.

Since the intervals of the variables ξ, η are defined by (10) and (9), the expected density of zeroes can be obtained.

$$\overline{\rho_0} = \lim_{\Delta t \to 0} \frac{\int_0^{\infty} \int_{\eta \Delta t}^{0} p(\xi, \eta) d\xi d\eta}{\Delta t} = \int_0^{\infty} \eta p(0, \eta) d\eta.$$

Similarly, for the interval between t and $t + \Delta t$, if the slope is negative or $-\infty < \eta < 0$, then

$$\overline{\rho_0} = -\int_{-\infty}^{0} \eta p(0, \eta) d\eta.$$

Combining these two expressions, (8) is obtained.

Relation Between ρ_0 and ξ^2, η^2

This relation is expressed by

$$\overline{\rho_0} = k_0 \sqrt{\frac{\overline{\eta^2}}{\overline{\xi^2}}} = k_0 \sqrt{\frac{\overline{f'(t)^2}}{\overline{f(t)^2}}}. \tag{11}$$

where k_0 is a constant and the bars indicate the ensemble average. For stationary time series, the time average and ensemble average are not distinguishable.

To prove (11), let the bivariate probability density function $p(\xi, \eta)$ be expressed in normalized form

$$p(\xi, \eta) = \xi_0^{-1} \eta_0^{-1} f\left(\frac{\xi}{\xi_0}, \frac{\eta}{\eta_0}\right) = \xi_0^{-1} \eta_0^{-1} f(x, y)$$

where

$$x \equiv \frac{\xi}{\xi_0} \qquad y \equiv \frac{\eta}{\eta_0}$$

and ξ_0 and η_0 are the bases of normalization.

In so doing, the two variables in the probability density function become dimensionless, and

$$p(\xi, \eta) d\xi d\eta = f(x, y) dx dy.$$

The ensemble averages of η^2 and ξ^2 are

$$\overline{\eta^2} = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \eta^2 p(\xi, \eta) d\xi d\eta$$
$$= \eta_0^2 \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} y^2 f(x, y) dx dy$$
$$= c_1 \eta_0^2 \tag{12}$$
$$\overline{\xi^2} = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \xi^2 p(\xi, \eta) d\xi d\eta$$
$$= \xi_0^2 \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} x^2 f(x, y) dx dy$$
$$= c_2 \xi_0^2. \tag{13}$$

The expected density of zeroes is, from (8),

$$\overline{\rho_0} = \int_{-\infty}^{\infty} |\eta| \, p(0, \eta) d\eta$$
$$= \eta_0 \int_{-\infty}^{\infty} |y| \, \xi_0^{-1}\eta_0^{-1} f(0, y) \eta_0 dy$$
$$= \frac{\eta_0}{\xi_0} \int_{-\infty}^{\infty} |y| \, f(0, y) dy$$
$$= c_3 \frac{\eta_0}{\xi_0} \tag{14}$$

where c_1, c_2, and c_3 are characteristic constants of the original distribution. Combining (12), (13), and (14)

$$\overline{\rho_0} = c_3 \sqrt{\frac{c_2}{c_1}} \sqrt{\frac{\overline{\eta^2}}{\overline{\xi^2}}} = k_0 \sqrt{\frac{\overline{\eta^2}}{\overline{\xi^2}}} = k_0 \sqrt{\frac{\overline{f'(t)^2}}{\overline{f(t)^2}}}$$

which is (11).

Thus, it is proved that for a stationary time series whose distribution is definite, as expressed by $p(\xi, \eta)$, the expected density of zeroes is proportional to the ratio of the root-mean-square value of the first time derivative and the root-mean-square value of the time series itself, provided the integrals contained in c_1, c_2, and c_3 also exist.

Relation Between $\overline{\rho_0}$ and the Autocorrelation Function

The autocorrelation function is defined as

$$\phi(\tau) = \overline{f(t) f(t - \tau)}.$$

The bar indicates either the time average or the ensemble average. For $\tau = 0$

$$\phi(0) = \overline{f(t)^2} = \overline{\xi^2}. \tag{15}$$

It may be proved that [20]

$$-\frac{d^2\phi}{d\tau^2} = -\phi''(\tau) = \overline{f'(t) f'(t - \tau)}.$$

For $\tau = 0$

$$-\phi''(0) = \overline{f'(t)^2} = \overline{\eta^2}. \tag{16}$$

Combining (11), (15), ane (16), the expected density of zeroes can be expressed alternatively as

$$\bar{\rho}_0 = k_0 \sqrt{\frac{\overline{f'(t)^2}}{\overline{f(t)^2}}} = k_0 \sqrt{\frac{-\phi''(0)}{\phi(0)}}. \tag{17}$$

It is proportional to the square root of the ratio of the negative second derivative of the autocorrelation function to the autocorrelation function itself both evaluated at $\tau = 0$.

Relation Between $\overline{\rho_0}$ and the Moments of the Power Spectrum

The power spectral density, $\Phi(f)$, is related to the autocorrelation function by the Fourier transform

$$\phi(\tau) = \int_0^{\infty} \Phi(f) \cos 2\pi f \tau df.$$

For $\tau = 0$,

$$\phi(0) = \int_0^{\infty} \Phi(f) df = M_0. \tag{18}$$

Similarly, it may be proved that

$$-\phi''(0) = 4\pi^2 \int_0^{\infty} f^2 \Phi(f) df = 4\pi^2 M_2. \tag{19}$$

Where M_0 and M_2 are used to designate the zero- and second-order moments of the power spectrum, $\Phi(f)$ versus f.

Combining (17), (18), and (19)

$$\bar{\rho}_0 = k_0 \sqrt{\frac{\overline{f'(t)^2}}{\overline{f(t)^2}}} = k_0 \sqrt{\frac{-\phi''(0)}{\phi(0)}} = 2\pi k_0 \sqrt{\frac{M_2}{M_0}}. \tag{20}$$

Equation (20) gives another interpretation to $\overline{\rho_0}$. It indicates that the expected density of zeroes is proportional to the square root of the ratio of the second moment ot the zeroth moment of the power spectrum.

The Proportionality Constant k_0 for Random Noise

The above derivations are specially applicable to the case when $f(t)$ represents random noise for which $p(\xi, \eta)$ is "Gaussian." To show this, consider the Fourier series representation of random noise

$$f(t) = \sum_{n=1}^{N} c_n \cos (\omega_n t - \phi_n). \tag{21}$$

Where $\phi_1, \phi_2 \ldots \phi_N$ are angles distributed at random over the range $(0, 2\pi)$

$$c_n = [2\Phi(f_n)\Delta f]^{1/2}$$
$$\omega_n = 2\pi f_n$$
$$f_n = n\Delta f.$$

Under certain conditions the "central-limit theorem" of probability can be applied and the distribution approaches the "normal" or "Gaussian" form. These

conditions may be stated qualitatively along with (21) as follows:[23]

(a) $f(t)$ has a large number of Fourier series components, i.e., $N\rightarrow\infty$.

(b) The ratio of the quadratic constant (i.e., the average power) of any one component in comparison with the total is vanishingly small, i.e., $c_n\rightarrow 0$.

(c) The phases of the components are distributed at random, i.e., ϕ_n a random variable.

(d) There is no direct-current component, i.e., $c_0=0$.

These conditions are usually satisfied by random noise. The probability density function $p(\xi, \eta)$ can be written in the form

$$p(\xi, \eta) = \frac{(\overline{\xi^2}\overline{\eta^2})^{-1/2}}{2\pi} \exp\left[-\frac{\xi^2}{2\overline{\xi^2}} - \frac{\eta^2}{2\overline{\eta^2}}\right]. \qquad (22)$$

Instead of going through the steps to derive (11) which is for the more general case, it is simpler here to substitute (22) directly into (8) whence

$$\overline{\rho_0} = \int_{-\infty}^{\infty} |\eta| \frac{(\overline{\xi^2}\overline{\eta^2})^{-1/2}}{2\pi} \exp\left[\frac{-\eta^2}{2\overline{\eta^2}}\right] d\eta = \frac{1}{\pi}\sqrt{\frac{\overline{\eta^2}}{\overline{\xi^2}}} \qquad (23)$$

[23] S. Goldman, "Frequency Analysis, Modulation and Noise," McGraw-Hill Book Co., New York, N. Y., p. 329; 1948.

Comparing (23) with (11), it is found that for a Gaussian distribution, the proportionality constant

$$k_0 = \frac{1}{\pi}.$$

For other distributions, k_0 would be different. It is interesting to note at this point that for a single sine wave of frequency f

$$f(t) = A \sin 2\pi ft.$$

If $\overline{\rho_0}$ is taken as $2f$, the proportionality constant between $\overline{\rho_0}$ and $\sqrt{\overline{f'(t)^2}/\overline{f(t)^2}}$ is also $1/\pi$. This perhaps suggests that for distributions other than Gaussian, as long as (5) applies, the proportionality constant k_0 may not be very much different from $1/\pi$.

The Expected Number of Maxima and Minima

By a similar process,[20] the expected number of maxima in a random function may be expressed as

$$\rho_m = k_m \sqrt{\frac{\overline{f''(t)^2}}{\overline{f'(t)^2}}} = k_m \sqrt{\frac{\phi^{IV}(0)}{-\phi''(0)}} = 2\pi k_m \sqrt{\frac{M_4}{M_2}} \qquad (24)$$

where k_m is a constant dependent upon the distribution, $\phi^{IV}(0)$ is the fourth derivative of the autocorrelation curve, $\phi(\tau)$ versus τ, and M_4 is the fourth moment of the power spectrum of $f(t)$.

Periodic-Waveguide Traveling-Wave Amplifier for Medium Powers*

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Summary—**A theoretical and experimental study of singly corrugated coaxial transmission lines is given here. The properties of the structure as a transmission line are calculated and the effect of the electron beam is taken into account by a field method. Theoretical values of gain and bandwidth are obtained. The results of the experimental study are compared with the theory. An amplifier giving 50-watts output with 20-db gain and 100-Mc bandwidth at a wavelength of 6.5 cm has been obtained. The best power output and efficiency which have been obtained are 125 watts and 7 per cent, respectively.**

Introduction

THEORETICAL ANALYSES and experimental studies of helix-type traveling-wave tubes have been carried out by Kompfner[1] and Pierce.[2] Other workers have contributed to the theoretical problem, notably Chu[3] who obtained a boundary value solution to the problem using Pierce's ideal helical current-sheet boundary and including the effect of Pierce's electron beam by the method of Hahn and Ramo. Field[4] has investigated other types of waveguide structures.

This paper is concerned with a traveling-wave tube using a corrugated inner-conductor coaxial transmission line with an annular electron beam between the inner and outer conductors. The first part deals with the theoretical analysis of the structure following the method of Goldstein[5] for a field solution in the absence of the electron beam and includes the electronic parameters by the method Chu used for the helix traveling-wave tube. The second part deals with an experimental

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‡ Federal Telecommunication Laboratories, Inc., Nutley, N. J.

[1] R. Kompfner, "Traveling wave valve," *Wireless World*, vol. 52, pp. 369–372; November, 1946.

[2] J. R. Pierce, "Theory of the beam-type traveling-wave tube," PROC. I.R.E., vol. 35, pp. 111–123; February, 1947.

[3] L. J. Chu and J. D. Jackson, "Field theory of traveling-wave tubes," PROC. I.R.E., vol. 36, pp. 853–862; July, 1948.

[4] L. M. Field, "Some slow-wave structures for traveling-wave tubes," PROC. I.R.E., vol. 37, pp. 34–40; January, 1949.

[5] H. H. Goldstein, "Cavity Resonators and Waveguides Containing Periodic Elements," Doctoral Thesis, Massachusetts Institute of Technology, 1943.

study and the comparison of the results with the theory.

The choice of the corrugated coaxial structure from the plethora of possible slow waveguide structures was made after a wide investigation. The type of amplifier described by Field and the iris-loaded waveguide structure proved to have very limited bandwidth; the singly corrugated coaxial line represented a compromise between gain, power output, and bandwidth.

Part I

A. Calculation of the Cold Phase Velocity

The structure we are considering, which is assumed to be lossless, is shown in Fig. 1. It belongs to a large class of periodic waveguides along with the linear magnetron and the iris-loaded waveguide used in linear-accelerator

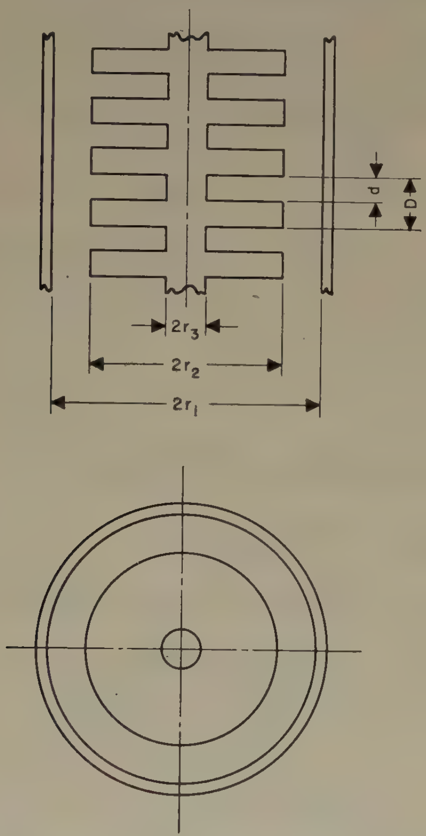

Fig. 1—Singly corrugated coaxial transmission line.

work. The properties of such structures are dealt with extensively by Brillouin,[6] Slater,[7] and Goldstein;[5] the latter treats in detail, but with certain approximations, the properties of our structure.

The method of solution is to obtain in the annular or "hole" region $r_1>r>r_2$ solutions satisfying Maxwell's equation and meeting the boundary conditions at $r=r_1$, and to obtain standing-wave solutions for the fields in the slots $r_2>r>r_3$, including higher-order cutoff modes. The field expansions are Fourier series, and by equating coefficients, an equation may be obtained with the required degree of approximation depending on the number of terms used.

The wave functions for circularly symmetric TM waves in this geometry are Bessel functions of order zero of the first and second kinds. The wave equation is easily separated and the following functions are obtained, taking into account that $E_z=0$ at $r=r_1$ and at $r=r_3$. For $r_1>r>r_2$

$$E_z = \sum_{\infty}^{\infty} A_n Z_0(\gamma_n, r, r_1) e^{ih_n z}$$
$$H_\theta = \sum_{\infty}^{\infty} \frac{ik^2 A_n}{\mu\omega\gamma_a} Z_0'(\gamma_n, r, r_1) e^{ih_n z} \tag{1}$$

where

$$k^2 = \gamma_n^2 + h_n^2 \quad \text{and} \quad h_n = H_0 + \frac{2\pi n}{D}\cdot$$

$$Z_0(\gamma_n, r, r_1) \equiv J_0(\gamma_n r) Y_0(\gamma_n r_1) - J_0(\gamma_n r_1) Y_0(\gamma_n r)$$

where the derivatives are taken with respect to the arguments, and k is the free-space wave number.

For $r_3 \leqq r \leqq r_2$ in the sth slot the fields can be written

$$E_z = \sum_{-\infty}^{\infty} e^{iH_0 sD}\left(a_p Z_0(k_p, r, r_3)\cos\frac{2\pi pz}{d} + b_p Z_0(K_p, r, r_3)\sin\frac{(2p+1)\pi z}{d}\right)$$

$$H_\theta = \sum_{-\infty}^{\infty} e^{iH_0 sD} ik^2\left(\frac{a_p}{k_p} Z_0'(k_p, r, r_3)\cos\frac{2\pi pz}{d} + \frac{b_p}{K_p} Z_0'(K_p, r, r_3)\sin\frac{(2p+1)\pi z}{d}\right) \tag{2}$$

where

$$k_p = \sqrt{k^2 - \left(\frac{2\pi p}{d}^2\right)}$$

and

$$K_p = \sqrt{k^2 - \left(\frac{(2p+1)\pi}{d}\right)^2}.$$

For one period of the structure at $r=r_2$ the longitudinal electric field must be zero for

$$\frac{D}{2} \geqq Z \geqq \frac{d}{2}$$

and for $|Z| \leqq d/2$, continuity of the fields is required. By the usual manipulation of the Fourier coefficients, (3) is obtained.

$$A_n Z_0(\gamma_n, r_2, r_1) D = \sum_p \sum_q \frac{A_q}{\gamma_q} Z_0'(\gamma_q, r_2, r_3) \frac{2}{d} R_{pqn}$$

[6] L. Brillouin, "Wave Propagation in Periodic Structures," McGraw-Hill Book Co., New York, N. Y.; 1946.
[7] J. C. Slater, "Electromagnetic Waves in Iris-Loaded Waveguides," Report #48, Massachusetts Institute of Technology, Research Laboratory of Electronics, Cambridge, Mass.; September 19, 1947.

$$R_{pqn} = \left\{ k_p \frac{4h_n h_q \sin\dfrac{h_n d}{2} \sin\dfrac{h_q d}{2}}{\left(\left(\dfrac{2\pi p}{d}\right)^2 - h_n^2\right)\left(\left(\dfrac{2\pi p}{d}\right)^2 - h_q^2\right)} \frac{Z_0(k_p, r_2, r_3)}{Z_0'(k_p, r_2, r_3)} \frac{1}{\epsilon_p} - K_p \frac{4h_n h_q \cos\dfrac{h_n d}{2} \cos\dfrac{h_q d}{2}}{\left(\left[(2p+1)\dfrac{\pi}{d}\right]^2 - h_n^2\right)\left(\left[(2p+1)\dfrac{\pi}{d}\right]^2 - h_q^2\right)} \frac{Z_0(K_p, r_2, r_3)}{Z_0'(K_p, r_2, r_3)} \right\} \tag{3}$$

where

$$\epsilon_p = 1 \qquad p \neq 0$$
$$\epsilon_p = 2 \qquad p = 0.$$

Equation (3) is a set of n homogeneous equations in n unknowns; it is the usual type of equation obtained in these problems and can be solved in principle. Provided the guide wavelength is large compared to D, the amplitudes of the space harmonics are small and one can consider the case where $n = p = q = 0$. This amounts to taking the principal mode in the hole and the first even and odd modes in the slot. For values of $H_0 \rightarrow (\pi/D)$, the space harmonics must be included and Slater[6] has given a more general method of solution to this type of problem.

We are concerned only with geometrics where $D < r_2$ and $|H_0 D/2\pi| < 1$. Under these conditions, letting $\gamma_0 = i\gamma_0$, and going to the modified Bessel functions, (3) becomes

right by equating at $r = r_2$ the values of E_z/H_θ averaged over a period of the structure for the fields of the two regions. This latter method does not allow one to calculate the effect of the odd slot mode, which gives an important correction to the phase velocity obtained from the simple method. A qualitative picture of the electric field is shown in Fig. 2.

For very low frequencies the asymptotic values of the Bessel functions for small arguments may be used in (4), except in the second term on the right, which is, for most traveling-wave tube geometries, very nearly unity. The resulting asymptotic phase velocity for low frequencies is

$$\frac{V}{C} = \frac{1}{\sqrt{\dfrac{\dfrac{8d^2}{\pi^3 D r_2} - \dfrac{d}{D}\log\dfrac{r_2}{r_3}}{\dfrac{8d^2}{\pi^3 D r_2} - \log\dfrac{r_1}{r_2}}}} \tag{5}$$

$$\gamma_0 \frac{I_0(\gamma_0 r_2)K_0(\gamma_0 r_1) - I_0(\gamma_0 r_1)K_0(\gamma_0 r_2)}{I_1(\gamma_0 r_2)K_0(\gamma_0 r_1) + I_0(\gamma_0 r_1)K_1(\gamma_0 r_2)} = \frac{d}{D} k \frac{J_0(kr_2)Y_0(kr_3) - J_0(kr_3)Y_0(kr_2)}{J_1(kr_2)Y_0(kr_3) - J_0(kr_3)Y_1(kr_2)} \left[\frac{\sin\dfrac{H_0 d}{2}}{\dfrac{H_0 d}{2}}\right]^2$$
$$+ \frac{8(H_0 d)^2}{D\pi^3}\left(\cos^2\frac{H_0 d}{2}\right) \frac{I_0\left(\dfrac{\pi r_2}{d}\right)K_0\left(\dfrac{\pi r_3}{d}\right) - I_0\left(\dfrac{\pi r_3}{d}\right)K_0\left(\dfrac{\pi r_2}{d}\right)}{I_1\left(\dfrac{\pi r_2}{d}\right)K_0\left(\dfrac{\pi r_3}{d}\right) + I_0\left(\dfrac{\pi r_3}{d}\right)K_1\left(\dfrac{\pi r_2}{d}\right)}. \tag{4}$$

It is convenient to regard (4) as an impedance equation. It can be obtained without the second term on the

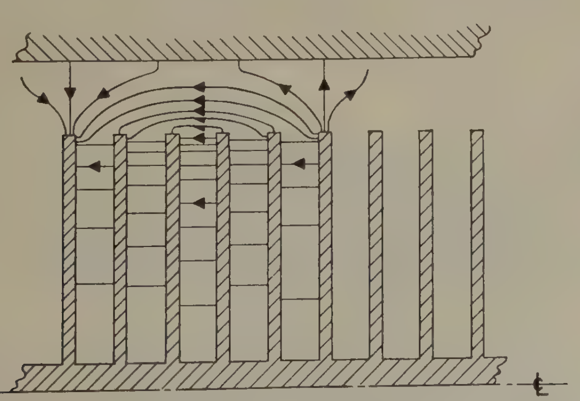

Fig. 2—Qualitative picture of the electric field.

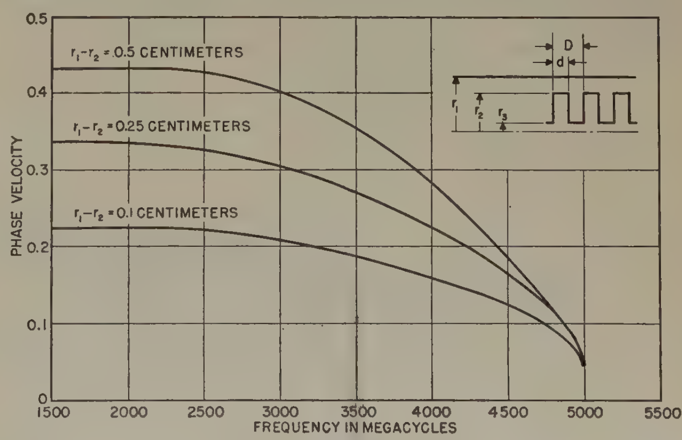

Fig. 3—Approximate theoretical phase velocity for a tube in which $d/D = 0.135$ and which has dimensions in centimeters of $r_2 = 1.2$, $r_3 = 0.4$, and $D = 0.15$.

which is independent of the frequency and always less than the velocity of light. Fig. 3 shows the results of an approximate phase-velocity calculation for a typical structure. Fig. 4 shows a comparison between measured results and the data computed from (3). The mean error is about four per cent and cannot be much reduced without including the effect of several space harmonics.

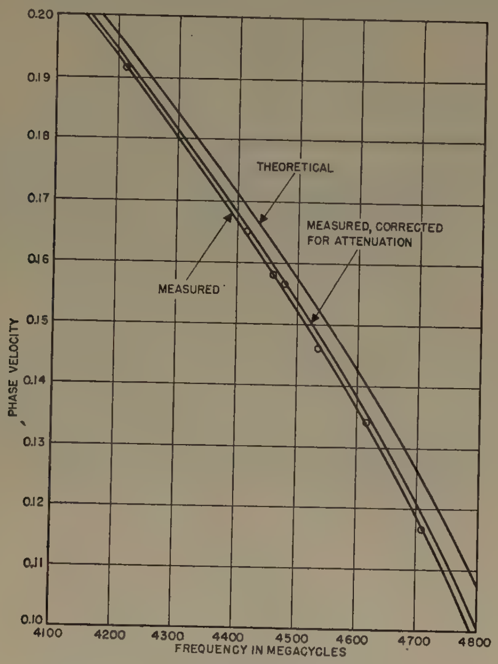

Fig. 4—Comparison of the experimental phase velocity with theoretical values calculated from the complete theory.

B. Properties with an Electron Beam Filling the Region, $r_2 \leqq r \leqq r_1$

We calculate here, by a small signal approximation, the modification of the cold properties introduced by a monochromatic electron beam constrained to move axially at a velocity small compared to that of light Under these conditions only the z component of the electric field (always neglecting the Lorentz forces) interacts with the electron stream, and it may easily be shown[3] that the ac current density has only a z component, which is linearly related to the longitudinal electric field by the relationship

$$J = \left(\frac{-\omega e/mJ_0}{v_0^3(\beta - H)^2}\right)E_z \text{ where } \beta = \frac{\omega}{v_0}.$$

With this relation the analysis of Section I*A* may be carried out and an equation analogous to (3) obtained.

$$\frac{\gamma}{1 - \dfrac{\alpha}{(\beta - H)^2}} P(\gamma, r_2, r_1) = \frac{d}{D} k_0 \frac{Z_0(k_0, r_2, r_3)}{Z_0'(k_0, r_2, r_3)} \left[\frac{\sin Hd/2}{\dfrac{Hd}{2}}\right]^2 + \left(\cos^2 \frac{Hd}{2}\right)\left(\frac{8(Hd)^2}{D\pi^3}\right) \tag{6}$$

where

$$P = \frac{I_0(\gamma r_2)K_0(\gamma r_1) - I_0(\gamma r_1)K_0(\gamma r_2)}{I_1(\gamma r_2)K_0(\gamma r_1) + I_0(\gamma r_1)K_1(\gamma r_2)}$$

and

$$\gamma^2 = (H^2 - k_0^2)\left(1 - \frac{\alpha}{(\beta - H)^2}\right)$$

and

$$\alpha = \frac{4\pi(e/m)\rho_0}{v_0^2} = 9.5 \times 10^4 \frac{J_0}{V_0^{3/2}} \frac{\text{amp/cm}^2}{\text{Volts}}.$$

Equation (6) can be solved exactly numerically, but it is far easier to obtain a solution by considering the hot propagation constant as a perturbation on the cold case. This method is in contradistinction to the method of Pierce[2] who considers the hot modes to be perturbations of the Hahn-Ramo waves traveling on the electron stream.

The propagation constant H, with the electron beam, differs from the cold constant H_0 by a small quantity δ which may be complex. One may modify (6) by expanding P in a Taylor series about $\gamma = \gamma_0$ to the first order in α and δ. Setting $H = H_0$ in the terms on the right (since the term is a correction, the error so introduced is negligible), the equation can be reduced to the following form:

$$\delta(u - \delta)^2 + A = 0 \tag{7}$$

where

$$u = \beta - H_0$$

and

$$A = \left\{\frac{\dfrac{\alpha}{2}\left\{\dfrac{d}{D} k \dfrac{Z_0(k, r_2, r_3)}{Z_0'(k, r_2, r_3)}\left[\dfrac{\sin \dfrac{H_0 d}{2}}{\dfrac{H_0 d}{2}}\right]^2 + \left(\cos^2 \dfrac{H_0 d}{2}\right)\left(\dfrac{8(H_0 d)^2}{D\pi^3}\right)\right\}}{\left(P + H_0 \dfrac{dP}{d\gamma}\right)}\right\}_{\gamma=\gamma_0}.$$

Equation (7) is a form of the usual cubic equation obtained in traveling-wave tube theory.

The imaginary part of the complex pair of roots is the gain in nepers per cm. Reduced values of the imaginary part of the complex pair of roots are shown in Fig. 5 and for the real part in Fig. 6. For calculating the frequency dependence of the gain, one must calculate A as a function of the frequency from the cold theory and then one readily gets the gain and hot phase velocity from Figs. 5 and 6 and (7).

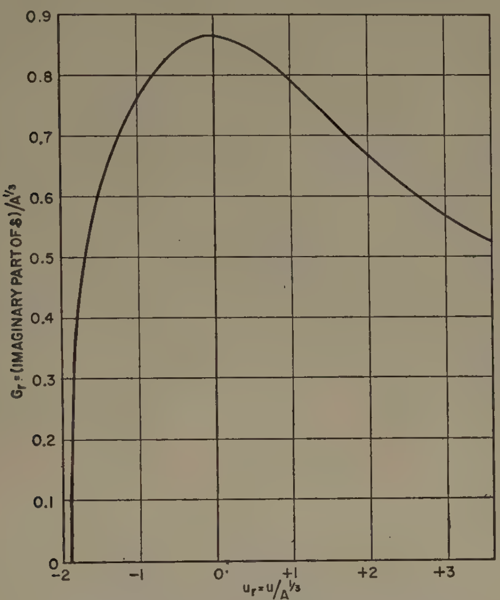

Fig. 5—Reduced imaginary part of the complex pair of roots of (7).

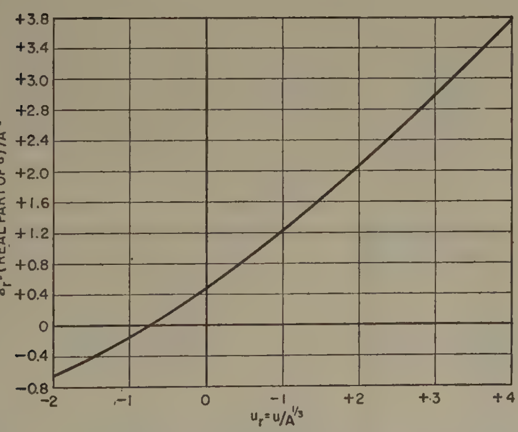

Fig. 6—Reduced real part of complex pair of roots of (7).

For small values of α, i.e., low beam current, the frequency dependence of the gain is mainly determined by the rate of change of u with frequency, i.e., the cold dispersion of the structure. For large values of α the frequency dependence of A becomes important.

Actual calculation of the performance of a given length of this structure depends on the initial boundary conditions and corrections for the presence of loss. The method in which these calculations are carried out is similar to that used for helix-type traveling-wave tubes for which extensive analyses are available in the literature.[2,3,8,9]

II. Experimental Study

A. Description of the Experimental Tube

A number of different structures based on the previous calculation have been made and tested. A drawing of a typical structure is shown in Fig. 7.

The corrugated inner conductor is made by assembling punched metallic disks with small metal spacers on a refractory metal mandrel and clamping the ends. The dimensions of the active part of the tube are, in the notation of Fig. 1,

$r_1 = 1.35$ cm	$d = 0.10$ cm
$r_2 = 1.20$ cm	$D = 0.125$ cm
$r_3 = 0.20$ cm	length $= l = 14.2$ cm.

The input and output of the corrugated coaxial structure are matched to 5/8-inch outside-diameter 50-ohm coaxial transmission line. The outer conductor has three longitudinal slots cut through it; the slots have been found to prevent propagation of an asymmetrical mode, which otherwise propagates freely and causes the amplifier to oscillate.

The annular electron beam is obtained from a pure tantalum ring emitter operated temperature limited for control of the beam current. The electrostatic and mag-

[8] M. Goudet, "Les Recents Progres des Tube Amplificateurs pour Ondes Centrimetriques," *Ann. Télécommun.*, vol. 3, pp. 445–455; December, 1948.

[9] O. E. H. Rydbeck, "Theory of Traveling-Wave Tubes," Ericsson Techniques No. 46, Telefonaktiebologet L. M. Ericsson, Stockholm; 1948.

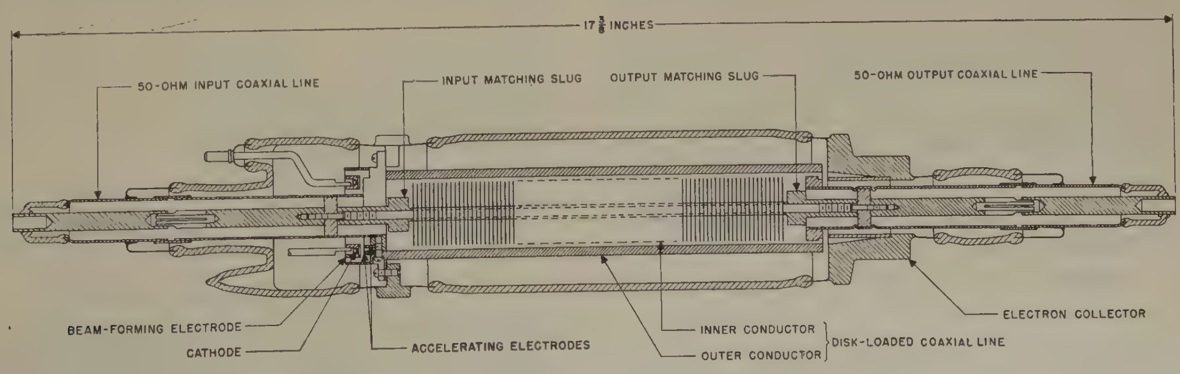

Fig. 7—Experimental tube.

netic fields are so controlled that the electron orbits are very nearly rectilinear and space charge forces, in the range of currents used, are generally small. The annular beam enters the coaxial structure through an annulus with four transverse spokes which serve to connect the outer conductors for radio-frequency currents. The beam leaves through a similar structure and is collected on a separate collector. The ratio of collected current to cathode current has varied between 0.6 and 0.8. The maximum possible beam efficiency is about 0.8 due to the part of the area of the annulus subtended by the radial spokes.

B. *Experimental Results and Comparison with Theory*

The operation of a dispersive traveling-wave tube of this type is characterized by a resonant dependence of the electronic gain and power output on frequency. The terminal impedance match is sufficiently broad-band so that the electronic properties alone determine the frequency response of the amplifier.

Fig. 8 shows the resonance voltage[10] as a function of frequency taken with very small beam currents. This curve gives a relation between beam voltage and frequency which allows one to plot the various parameters as a function of this beam voltage. Operation in a region of power saturation causes an upward shift in the resonance beam voltage, but the shift has never exceeded 10 per cent.

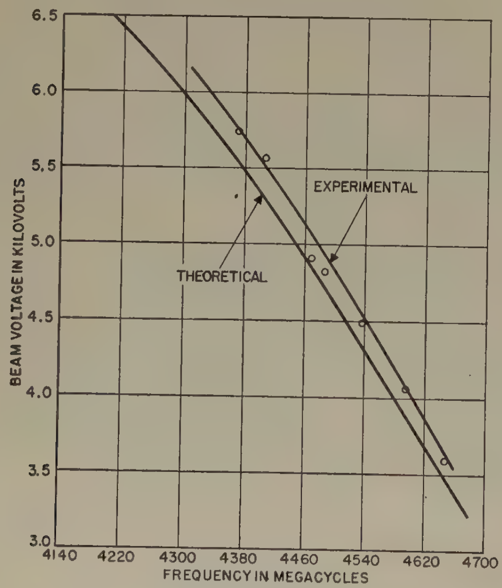

Fig. 8—Experimental resonance beam voltage for very small beam currents.

Fig. 9 shows the small-signal gain of the tube, with finite cold attenuation, as a function of beam current for various values of the beam voltage, which corresponds to the values of frequency given in parenthesis.

The gain of the amplifier as a function of power level exhibits the usual marked saturation effect seen in traveling-wave tubes. As the power input increases, a point is finally reached where an increase in the input power results in a decrease in the output power. The power output is then the maximum power output of the tube.

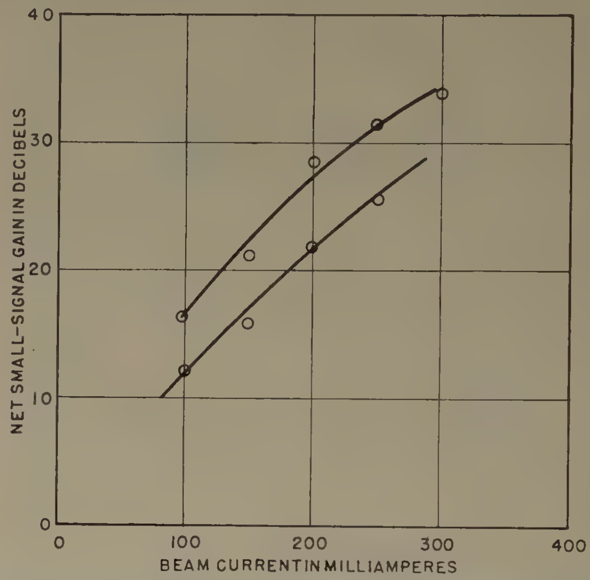

Fig. 9—Small-signal gain plotted against beam current. The values of V_0 in kilovolts and frequency in megacycles for the upper curve are 3.6 and 4,650 and, for the lower curve, 4.6 and 4,550.

Fig. 10 shows the saturation power output as a function of beam current. In the curves in Fig. 10 a slight shift towards higher voltage has taken place due to saturation.

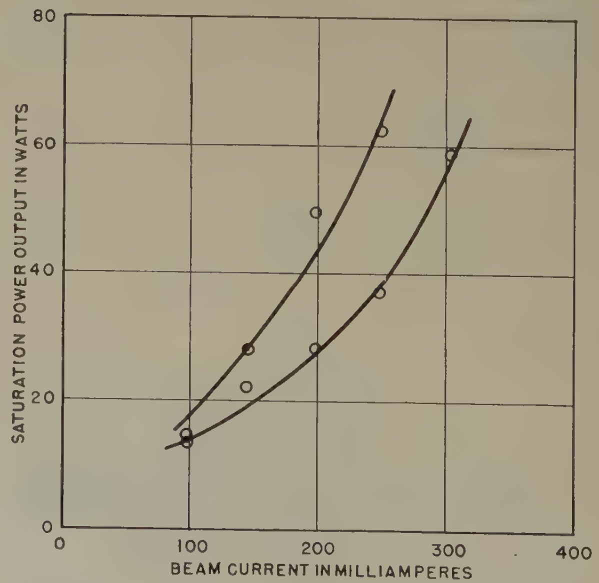

Fig. 10—Saturation power output plotted against beam current. In the upper curve, $V_0=4.9$ kv, $f=4{,}550$ Mc, and the gain is 16.9 db. The values for the lower curve are, respectively, 4.0, 4,650, and 20.

[10] The ratio of d/D for the theoretical curve is larger than that for the experimental curve which explains the higher experimental voltages in contrast to Fig. 4.

In comparing these results with the theory, it is necessary to take into account the excitation of the three forward waves at the input and the effect of the insertion loss. Excitation of the three forward waves results in a loss calculated[2,11] to be 9 db in the growing wave at synchronism. The value of this excitation loss can be shown by (7) to depend in a complicated fashion on the cold attenuation and the separation of the beam velocity from that for synchronism. The nature of this dependence is not sensitive and the value of excitation loss varies from 6 to 12 db for practical ranges of operation. The correction obtained for attenuation L, uniform along the structure, has been calculated by numerous authors[2,8] and it can be shown that (7) leads to the same conclusion, that one must subtract one third the attenuation from the gross electronic gain for small values of attenuation. The behavior for values of attenuation comparable to the gain can be shown by (7) to require subtracting about one half the total loss. Subject to the use of approximate values, the formula relating the net gain G_{net} to the gain per unit length G which was calculated in Part I is

$$G_{\text{net}} = lG - 9 - \frac{2L}{5} \text{ db.} \tag{8}$$

The values G in db per cm obtained from the experimental results by (8) are compared to the theoretical values in Fig. 11.

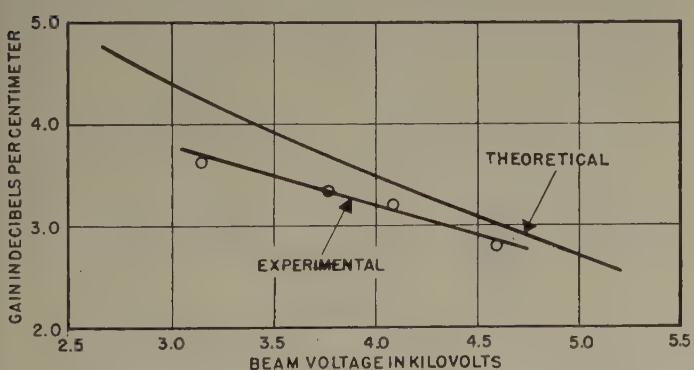

Fig. 11—Gain per unit length for an experimental tube with small-signal input compared with theoretical values, for 250 ma beam current.

The frequency dependence of the gain for fixed beam voltage is readily calculated from (7), and Fig. 12 shows the experimental gain as a function of frequency compared to the theoretical value, again by means of (8). The tube was operating partly saturated when this curve was taken, which has resulted in a shift of the center frequency of the experimental curve towards the theoretical value.

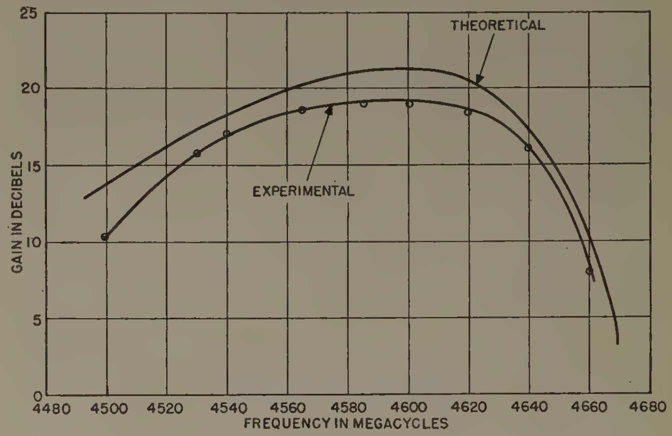

Fig. 12—Experimental and theoretical gain plotted against frequency. In the experimental case, the power input was 0.50 watt. The beam values were 4 kv and 200 ma.

C. Power Output

The power output of a traveling-wave tube cannot be calculated from the foregoing analysis due to the failure of the small signal approximation. In general, numerical techniques are required. However, experimental studies of helix tubes and qualitative theoretical arguments indicate that the power output is in general given by

$$P_{\text{sat}} = \frac{KG\lambda_g}{8.68\pi\sqrt{3}} I_0 V_0 \tag{9}$$

where G is the electronic gain in db per cm and K is a constant which depends on a variety of factors; K is most sensitive to the uniformity and losslessness of the output section of the tube. For helix tubes, values of K as high as three have occasionally been obtained. We have been largely unable to control the magnitude of the attenuation in the output section of our tube. The values of K corresponding to the measured powers shown in Fig. 10 range between 0.9 and 1.2. The best values of efficiency and power output that have been obtained are about 7 per cent and 125 watts at about 4.5 kv and 380 ma beam current with a gain of about 17 db. In view of the work with the helix tubes and of certain theoretical arguments, it is likely that efficiencies the order of 15 to 25 per cent and power outputs in the vicinity of 500 watts are possible in this frequency range with tubes of this general type.

[11] J. R. Pierce, "Effect of passive modes in traveling-wave tubes," Proc. I.R.E., vol. 36, pp. 993–996; August, 1948.

Gain of Electromagnetic Horns*

W. C. JAKES, JR.†, ASSOCIATE, IRE

Summary—An experimental investigation of the gain of pyramidal electromagnetic horns is described. For the horns tested it was found that (1) the "edge effects" are less than 0.2 db so that the gain of the horns may be computed to that accuracy from their physical dimensions and Schelkunoff's curves; and (2) for the transmission of power between two horns the ordinary transmission formula is valid, provided that the separation distance between the horns is measured between the proper reference points on the horns, rather than between their apertures.

I. Introduction

THE CUSTOMARY method of measuring the gain of large microwave antennas is by comparison with a small standard pyramidal horn. The gain of the standard horn is usually determined by calculation from the physical dimensions of the horn and use of curves given by Schelkunoff.[1] Since these curves are based on the assumption that the aperture field of the horn is the same as though the sides were continued indefinitely, it is apparent that the computed gain of the horn may be somewhat in error because of the doubtful validity of this assumption.

An experimental determination of the amount of error in the theoretically calculated gain due to this "edge effect" could be made by measuring the power transmitted between two identical horns at a separation distance r, measured between apertures, great enough so that the familiar transmission formula holds:

$$P_R = \left(\frac{G\lambda}{4\pi r}\right)^2 P_T \tag{1}$$

where

P_R = received power
P_T = transmitted power
G = gain of each individual horn
λ = free-space wavelength.

Any measurable difference between the gain computed from the horn dimensions and that given by (1) may be ascribed to edge effects.

Several considerations complicate the simple experiment described above. Ordinarily it is not practicable to make transmission measurements at extremely large values of r where it is reasonably certain that (1) holds. One condition, at least, that must be fulfilled is that the variation in phase of the transmitted wave over the aperture of the receiving horn should not exceed $\lambda/16$. If the transmitter were a point source, this would fix the minimum separation distance r_{min} between the two antennas as

$$r_{min} = 2\frac{b^2}{\lambda}, \tag{2}$$

where b is the larger dimension of the horn aperture. Since the transmitting antenna is not a point source, there is some uncertainty about the point from which to measure r_{min}; however if one measures from the aperture plane of the transmitter horn, it seems reasonable that the phase error will not exceed $\lambda/16$ at an r_{min} given by (2). It is not necessarily true, however, that if r is the distance between aperture planes the transmission formula (1) will be obeyed for the entire range of $r_{min} < r < \infty$.

II. Experiments

The experimental part of this study was carried out at a wavelength of 1.25 cm, as the distances and physical dimensions of the horns involved become small and easily managed in this range. The variation of P_R with r (between apertures) for $40\lambda \leq r \leq 200\lambda$ was measured for a number of pyramidal horns of various dimensions. Fig. 1 shows the physical setup employed. To reduce the effect of reflections, no objects were allowed to come closer than 70λ to the center line of the horn.

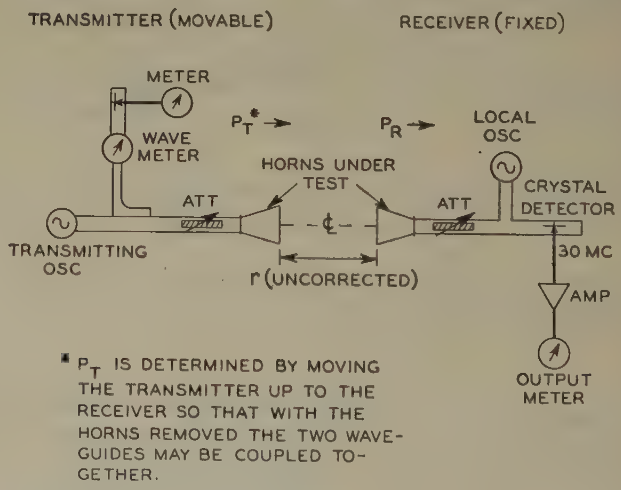

Fig. 1—Physical setup for measuring the variation with distance of the power transmission between two horn antennas.

Before listing the experimental results it will be helpful to give the horn nomenclature, as shown in Fig. 2. Note that in general the E-plane and H-plane slant lengths, l_E and l_H, are not necessarily equal. The "axial height" of the horn will be designated by h; if the horn

* Decimal classification: R325.82. Original manuscript received by the Institute, May 9, 1950; revised manuscript received, August 22, 1950.
† Bell Telephone Laboratories, Inc., Holmdel, N. J.
[1] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand, Inc., New York, N. Y., pp. 363–365; 1943.

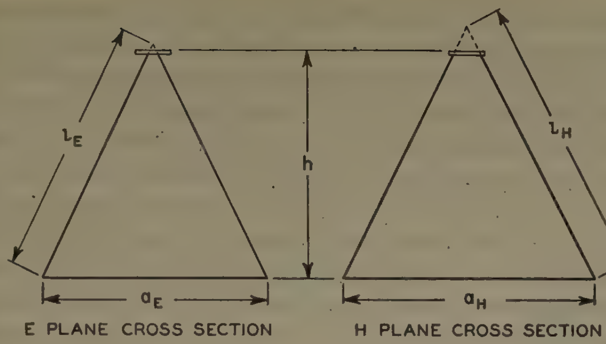

Fig. 2—Nomenclature for horn flared in both planes.

is optimum[2] the axial height will be designated by h_0.

In all, four pairs of horns were constructed and tested. They were made from sheet brass of 1/16 inch thickness; their physical dimensions are given in Table I. Note that horns 1 and 4 were optimum horns.

TABLE I

Horn	a_E	a_H	h
1	4.72λ	6.03λ	$10\lambda\ (=h_0)$
2	4.72λ	6.03λ	$20\lambda\ (>h_0)$
3	4.72λ	6.03λ	$5\lambda\ (<h_0)$
4	6.73λ	8.49λ	$20\lambda\ (=h_0)$

Curve A of Fig. 3 shows the experimental results for a pair of optimum horns (No. 1 in Table I); this is typical of the results obtained in general. It is to be noted that P_R does not vary as $1/r^2$, the departure being greater as r decreases. A distance d was found which, when added to the r co-ordinates of curve A, caused these points to lie on a straight line (curve B) whose slope corresponds to an inverse square variation of received power with distance. This indicates that if the separation distance is measured between the proper reference points on the horns, the inverse square relationship of (1) will be obeyed. The distance from the horn aperture back to this reference point will be called D.

Since the transmitting and receiving horns were identical in the above experiments, it follows that $D=d/2$. For the *optimum horns* (with $h=h_0$) D was found to be equal to the *axial height*. However, for the other horns the following was observed: if $h>h_0$, (horn 2) $D<h$; if $h<h_0$, (horn 3) $D>h$.

Since it has been experimentally demonstrated that (1) is valid for $r_{\min}<r<\infty$ provided r is measured between the proper horn reference points, this equation may now be used with the proper r to compute the actual horn gain. Curve B of Fig. 4 shows the results of this computation for horn 1. For comparison, curve A of Fig. 4 was computed using for r in (1) the separation distance between horn apertures. Curve C is the gain calculated from the physical dimensions of the horn

[2] An optimum horn is one for which the flare angles in both planes are so chosen that, for a given length of horn, the gain is a maximum. This follows if:

$$a_E^2 \doteq 2l_E\lambda;\ a_H^2 \doteq 3l_H\lambda.$$

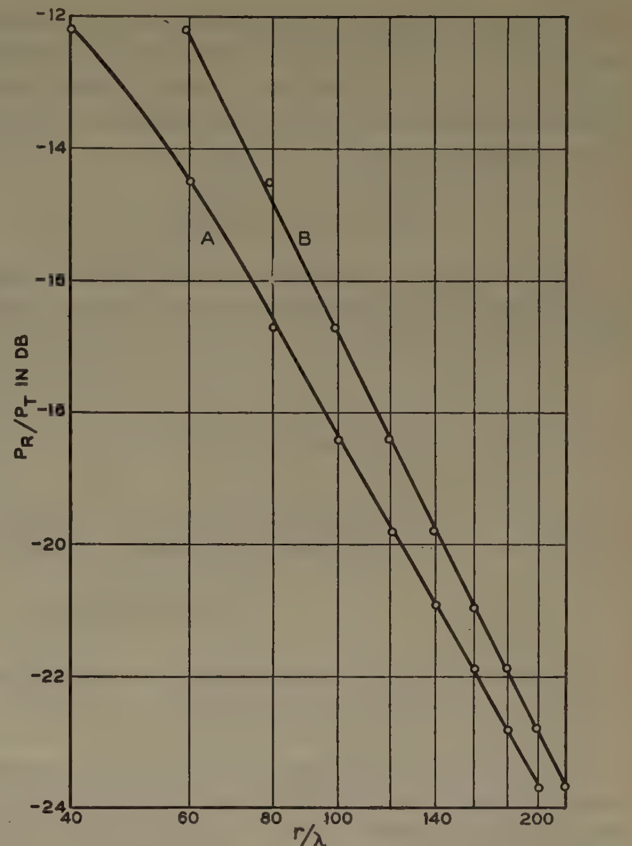

Fig. 3—Variation with distance of the power transmission between two horns. A. Experimental results. B. Experimental results corrected by adding a constant, d/λ, to the abscissae of the observed points.

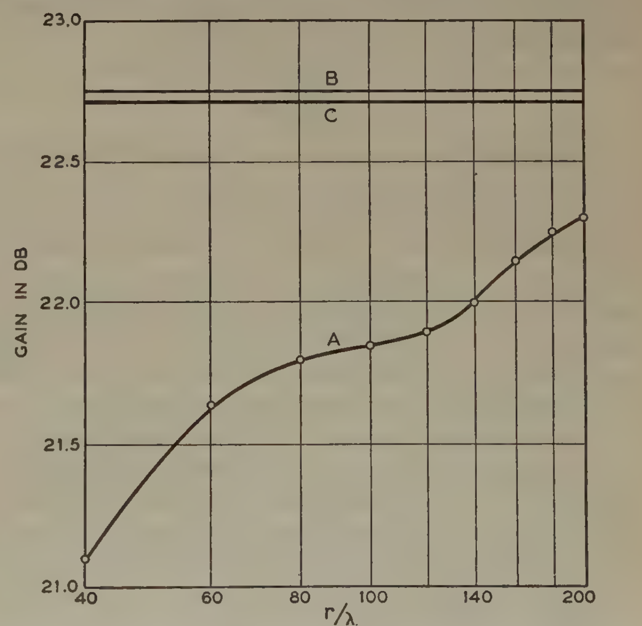

Fig. 4—Theoretical and experimentally observed horn gain. (Horn No. 1 in Table I.) A. Gain computed from (1) using for r the separation distance between apertures. B. Gain computed from (1) using for r the value r_0+d, where r_0 is the separation between apertures, d is a constant described in the text. C. Gain computed from the physical dimensions of the horn and Schelkunoff's curves.

and Schelkunoff's curves. These three curves are representative of the results for the four pairs of horns; in general, the difference between curves B and C did not exceed 0.2 db, and for the optimum horns it was less than 0.1 db.

III. Conclusions

When computing the gain of pyramidal electromagnetic horns it is permissible to use their actual physical dimensions and Schelkunoff's curves. The error due to edge effects is less than 0.1 db for optimum horns, with aperture dimensions greater than 4λ.

If it is desired to compute the transmitted power between two identical horns, (1) is valid even in the transition zone between the Fraunhofer and Fresnel regions, provided r is replaced by r_0+2D. Here, r_0 is the separation between apertures and D is described above.

D for an optimum horn is equal to the axial height, but for horns shorter or longer than optimum, D is greater or less than the axial height and must be determined by experiment.

Evaluation of Coaxial Slotted-Line Impedance Measurements*

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Summary—**Most ultra-high-frequency impedance measurements are made by detecting the voltage-standing-wave ratio and nodal position in a slotted section of coaxial transmission line. Sources of error in these measurements are discussed and methods of eliminating or evaluating them are presented. It is shown that the maximum error due to structural defects in determining the relative voltage can be predicted experimentally for most standing-wave machines and that the resulting maximum error in the voltage-standing-wave ratio is twice the maximum error in determining the relative voltage. The resulting maximum error in the nodal position and, also, the fractional errors in the load resistance and reactance due to the errors in the voltage-standing-wave ratio and nodal position are calculated and presented in graphical form.**

I. Introduction

As a part of a program to establish standards of measurement of electrical quantities at ultra-high frequencies, the accuracy of impedance measurements using current techniques has been evaluated. The accuracy of measurement of most other uhf electrical quantities is related to the accuracy of impedance measurements. The slotted section of transmission line with a traveling probe is the most widely used measuring instrument in this frequency range. This instrument utilizes what can be regarded as a comparison method, in which the impedance to be measured is compared with the characteristic impedance of the slotted section. Since the characteristic impedance can be determined accurately from the physical dimensions and the electrical properties of the slotted section, the instrument can be considered an "impedance meter" with an inherent reference standard.

In most uhf impedance measurements, the comparison of the unknown impedance with the characteristic impedance of the slotted section is made by determining the voltage-standing-wave ratio (VSWR) and a nodal position by either direct or indirect methods. Obviously, the accuracy of such impedance measurements is a function of the accuracy of the determination of VSWR and nodal position. The purpose of this paper is to present: (1) a brief survey of the literature on sources of error in measurements of VSWR and nodal position together with references to methods of eliminating or evaluating the errors; (2) a study of the maximum errors in determining VSWR and nodal position which result from errors in measuring relative-voltage distribution along the slotted section; and (3) curves for evaluating the errors in impedance measurements resulting from these errors in VSWR and nodal position.

II. Sources of Error in VSWR and Nodal Position Measurements

The principal errors in the determination of VSWR and nodal position originate in the voltage generator, detector, and slotted section. These errors are discussed in the following paragraphs, together with references which describe techniques for reducing the errors to a negligible amount or evaluating the magnitude of the errors.

1. Impure and Unstable UHF Voltage Input

The harmonic content of most unfiltered uhf power sources is of sufficient magnitude to cause serious errors in the measurement of both VSWR and nodal position,

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† Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C.

especially if an untuned probe is used. The probe is usually part of a high-Q circuit which is tuned to the fundamental frequency and discriminates against even harmonics. Low-pass filters with low insertion loss can be inserted between the generator and the slotted section to limit harmonic content to a negligible percentage of the fundamental.[1]

Instability and unwanted frequency modulation of the voltage generator output during an impedance measurement are other sources of error.[2] Variations in the amplitude of the uhf voltage generator output also cause errors in the measured values of VSWR. Fortunately, the amplitude and frequency stability of most uhf generators is sufficient for impedance measurements if the generators are operated from highly stabilized power supplies,[3] and proper care is taken to reduce unwanted frequency modulation.

2. *Undetermined Response Law of Detector System*

The voltage standing-wave existing in a slotted section is determined by sampling the electric field along the length of the slotted section and taking the ratio of the maximum to the minimum field. The response law of the detector system (consisting of detector element, amplifier unit, and indicator) can be determined by terminating the slotted section in a short circuit and plotting the indicated output as a function of probe position.[2,4] If the necessary equipment is available, it is preferable to determine the law of the detector system with a calibrated attenuator. The response of detector elements, such as Littelfuses, Wollaston wires, and barretters can be considered "square law" for low-power levels or for small values of VSWR. Of course, the accuracy of indication of any detector is limited by the accuracy of associated dc and af measuring instruments.

3. *Alteration of Standing-Wave Pattern by the Probe*

The error in measurement of VSWR and nodal position caused by the loading of the slotted section of transmission line by the probe is negligible if the probe is lossely coupled to the line. If the probe penetration does cause distortion in the observed standing-wave pattern, this distortion can be determined by noting the interaction between two identical probes in the same slotted section. If the motion of either probe causes a change in the indicated output of the other probe, which is kept stationary at various selected positions, then the probe penetration is considered excessive. If it is necessary to tightly couple the probe to the slotted section of transmission line in order to perform standing-wave measurements, the magnitude of the errors introduced in the measurement of VSWR and nodal position can be determined as a function of the effective admittance of the probe shunting the slotted section.[5-6]

4. *Attenuation*

Attenuation in a slotted section is caused by dielectric, conductor, and radiation losses. These losses are usually of such small magnitude that they seldom affect VSWR measurements. However, errors in impedance calculated from VSWR and nodal position measurements can occur if the attenuation between the load terminals and the position at which the VSWR is measured is large. This error can be corrected if the attenuation constant is known.[7]

5. *Effect of Slot*

A narrow longitudinal slot cut in the outer conductor of a section of coaxial transmission line decreases the capacitance per unit length of line[4] and permits radiation losses. The radiation losses are negligible for most types of measurements, and the capacitance decrease causes a slight increase in the characteristic impedance of the transmission line. The ends of the slot cause impedance discontinuities at the junction of the slotted and unslotted sections. The approximate residual VSWR introduced by these discontinuities can be easily calculated.[4] A commercially available $\frac{7}{8}$-inch standing-wave machine was found to have a calculated residual VSWR of less than 1.004 due to the slot. If the effect of discontinuities at the end of the slot and in connectors is appreciable, equivalent impedances can be determined experimentally.[8-12]

6. *Structural Defects of Slotted Section and Probe Carriage*

It may be concluded that the errors in the measurement of VSWR and nodal position discussed above can

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[2] Radio Research Laboratory Staff, "Very-High Frequency Techniques," McGraw-Hill Book Co., Inc., New York, N. Y., vol. 1, p. 35; 1947.

[3] R. C. Ellenwood and H. E. Sorrows, "Cathode heater compensation as applied to degenerative d.c. power supplies," *Jour. Res. Nat. Bur. Stand.*, vol. 42, no. 9; September, 1949.

[4] MIT Radiation Laboratory Series, "Techniques of Microwave Measurements," McGraw-Hill Book Co., Inc., New York, N. Y., vol. 11; 1947.

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[7] G. Glinski, "The solution of transmission-line problems in the case of an attenuating transmission line," *Trans. AIEE*, vol. 65, pp. 46–49; February, 1946.

[8] E. Feenberg, "The relation between nodal positions and standing-wave ratio in a composite transmission system," *Jour. Appl. Phys.*, vol. 17, pp. 530–532; June, 1946.

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[10] M. S. Wong, "Microwave impedance through an arbitrary four-terminal network," Presented, URSI-IRE Spring Meeting, Washington, D. C., 1948.

[11] D. M. Kerns, "The Basis of the Application of Network Equations to Waveguide Problems," CRPL-9-5, National Bureau of Standards; 1948.

[12] M. H. Oliver, "Discontinuities in concentric-line impedance-measuring apparatus," *Jour. IEE*, Part III, vol. 97, p. 29; January, 1950.

be evaluated or reduced to a negligible magnitude by the use of appropriate experimental techniques. It is, however, difficult if not impossible to evaluate or completely correct for the errors in VSWR and nodal-position measurements caused by structural defects or mechanical irregularities of the slotted section and probe carriage.[5] The sag of the center conductor between supports, the mechanical irregularities in probe carriage ways which result in vertical or lateral motion of the probe, and the nonuniformity of cross section of the coaxial slotted section are examples of structural defects. Mechanical irregularities in the probe carriage ways (slotted section mounted horizontally) which result in vertical or lateral displacement of the probe relative to the center conductor as the probe traverses the length of the slotted section are usually found to be the chief sources of error. Because of the large number of parameters involved, it is difficult to predict theoretically the errors in the measurement of VSWR for a given slotted section, but these errors can be predicted from experimental data. One method which can be used to predict the performance of a standing-wave machine involves approximately matching the load end of the slotted section in order that the irregularities in detector response may be measured as a function of probe position along the line. Experimental data for predicting the performance of a selected standing-wave machine are shown in Fig. 1. For this particular standing-wave

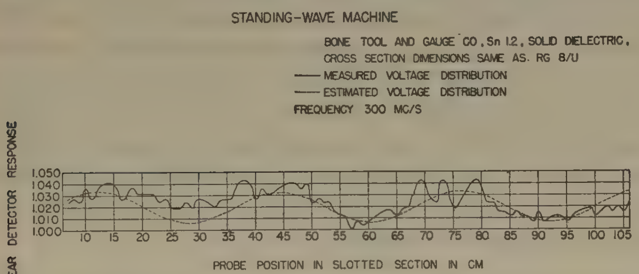


Fig. 1—Measured and estimated uhf voltage distribution for low VSWR.

machine there is a maximum deviation of the measured relative-voltage distribution of about ±2 per cent from the estimated, true relative-voltage distribution. The magnitude of this deviation was found to be constant for frequencies in the range of 300 to 600 Mc. It can, therefore, be assumed that there is a maximum error of 2 per cent in the determination of relative-voltage distribution in this standing-wave machine when measurements are made in the indicated frequency range.

Methods using either audio-frequency voltages or precise mechanical measurements for calibrating or determining the performance of a given slotted section are desired. However, though experimental studies on some slotted sections show correlation, on others no correlation is observed between af, mechanical, and uhf measurements. It is, therefore, concluded that the performance of a slotted-section type of standing-wave machine can be best determined by experiments performed at uhf. The important property to be determined is the fractional error in measuring the relative-voltage distribution. From this, the maximum fractional errors in measuring VSWR and nodal position can be determined, and from these, in turn, can be obtained the resulting errors in impedance measurements.

II. Accuracy of Measurement of VSWR and Nodal Position[13]

Theoretically, all errors in the measurements of VSWR and nodal position can be eliminated or evaluated except those errors caused by structural defects in the standing-wave machine. It has been shown that the fractional error (b) in the determination of relative voltage due to these structural defects can be obtained experimentally. The maximum error in the measured value of VSWR (ρ_0) resulting from this fractional error (b) can then be predicted. Since

$$\rho = \frac{V_{\max}}{V_{\min}}, \tag{1}$$

where ρ, $V_{\max}$, and $V_{\min}$ are, respectively, the true values of VSWR, voltage maximum, and voltage minimum, the limiting values of the measured VSWR (ρ_0) are

$$\frac{V_{\max}(1-b)}{V_{\min}(1+b)} \leqq \rho_0 \leqq \frac{V_{\max}(1+b)}{V_{\min}(1-b)}. \tag{2}$$

For small values of b, (2) reduces to

$$\rho(1-2b) \leqq \rho_0 \leqq \rho(1+2b). \tag{3}$$

Therefore, the maximum fractional error in measured VSWR will not be more than twice the maximum fractional error in relative-voltage measurement. The maximum error in measurement of VSWR with the standing-wave machine for which sample experimental curves are shown in Fig. 1 would thus be less than 4 per cent.

The error in the determination of the nodal position can also be related to the error in relative-voltage measurement. The nodal position is usually obtained by finding the midpoint between two positions of equal detector response on each side of the minimum. A method of determining the error in measured nodal position resulting from errors in relative-voltage measurements is best described with the aid of Fig. 2. The solid line represents the true voltage distribution (V) and the dashed lines represent the limits of the measured values of relative voltage, $(1 \pm b)V$. The measured values, therefore, lie within the area bounded by the dashed lines. There is seen to be a maximum error of $\pm\delta\theta$ in the location of a given voltage because of possible errors in relative-voltage measurements. An error in the location of the nodal position results from this error, $\delta\theta$, in location of the equal-response voltages.

[13] The analysis in Sections III and IV applies to waveguides as well as to coaxial transmission lines.

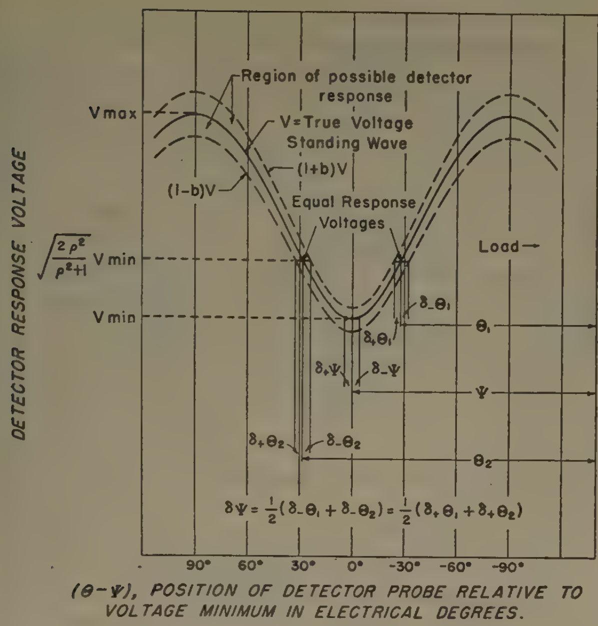


Fig. 2—The determination of the maximum error in nodal position from the fractional error b in relative-voltage measurement.

It is shown in Appendix A that, if it is assumed that the slotted section is lossless and that the probe does not distort the standing-wave pattern, the maximum error ($\delta\Psi$) in the location of the nodal position is given by the equation

$$\delta\Psi = \frac{1}{4}\arccos\left[\frac{\rho^2+1}{\rho^2-1} - \frac{1+(\rho^2-1)\sin^2(\theta-\Psi)}{\frac{(\rho^2-1)}{2}(1-b)^2}\right] - \frac{1}{4}\arccos\left[\frac{\rho^2+1}{\rho^2-1} - \frac{1+(\rho^2-1)\sin^2(\theta-\Psi)}{\frac{(\rho^2-1)}{2}(1+b)^2}\right], \tag{4}$$

where $(\theta-\Psi)$ is the magnitude of the distance in electrical degrees from the nodal position to the true equal-response voltage positions.

$\delta\Psi$ is seen to be a function of ρ, b, and $(\theta-\Psi)$. For a given impedance measurement, ρ is the measured VSWR and b is a constant for the slotted section used, whereas $(\theta-\Psi)$ is a function of the equal-response voltages.

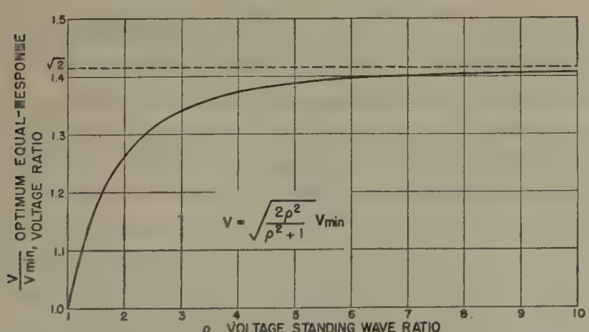


Fig. 3—Curve of optimum equal-response voltage for minimum error in determination of nodal position.

It is shown in Appendix B that $\delta\Psi$ is a minimum if equal-response voltages are selected having a magnitude

$$V = \sqrt{\frac{2\rho^2(1+b^2)}{\rho^2+1}}\,V_{\min}. \tag{5}$$

Since b^2 is very small compared to unity, it may be neglected. V rapidly approaches $\sqrt{2}\,V_{\min}$ as ρ increases as shown in Fig. 3. From equations (4) and (5), the minimum value, $\delta\Psi_m$, of the maximum error in nodal position is found to be

$$\delta\Psi_m = \frac{1}{4}\arccos\left[1 - \frac{32\rho^2b^2}{(\rho^2-1)^2(1-b^2)^2}\right]. \tag{6}$$

Curves for (6) are shown in Fig. 4.

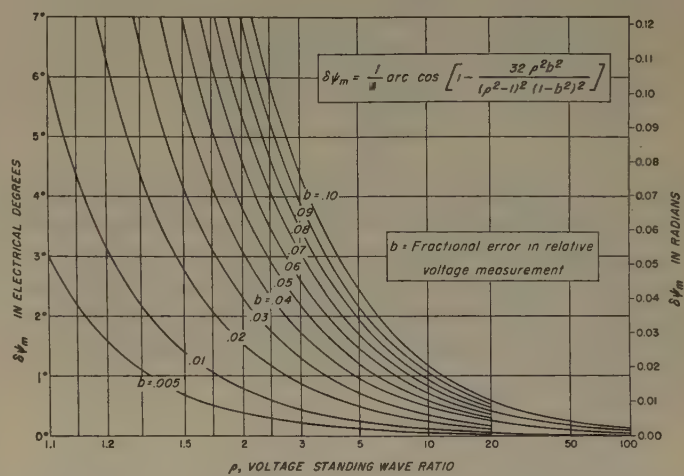


Fig. 4—Curves of maximum error in nodal position determined from positions of equal-response voltages of

$$\sqrt{\frac{2\rho^2(1+b^2)}{\rho^2+1}}\,V_{\min}$$

versus voltage-standing-wave ratio for various fractional errors b in measurement of relative voltage.

For small values of b and of the error $\delta\Psi_m$, (6) can be approximated by

$$\delta\Psi_m \simeq \frac{2\rho b}{\rho^2-1}. \tag{7}$$

It is easily shown that the VSWR and nodal-position measurements are meaningless for $\rho \leqq (1+b/1-b)$.

III. Evaluation of Errors in Impedance Measurements as a Function of the Error in VSWR and Nodal Position[14]

The terminating impedance of a lossless uniform transmission line can be expressed as a function of the

[14] Following completion of this paper it was brought to the authors' attention that an essentially identical development of the material presented in Section III and Figs. 5–8a was independently given by F. M. Millican in U. S. Navy Electronics Laboratory Report No. 110, "Error Analysis of Slotted Transmission Line Impedance Measurements."

VSWR (ρ) and the distance (Ψ) from the nodal position to the load terminals by the following equation:

$$Z = \frac{\cos\Psi - j\rho\sin\Psi}{\rho\cos\Psi - j\sin\Psi}, \tag{8}$$

where Z is the normalized impedance at the load terminals. The assumption, in (8), that the slotted section is lossless introduces only a negligible error in most measurements. However, if the section of transmission line between the load and the measured nodal position has excessive loss, (8) must be modified as described in the literature.[7]

The normalized resistive and reactive components of the complex impedance Z are given respectively by the following equations:

$$r = \frac{\rho}{\rho^2\cos^2\Psi + \sin^2\Psi} \tag{9}$$

$$x = \frac{(1-\rho^2)\sin\Psi\cos\Psi}{\rho^2\cos^2\Psi + \sin^2\Psi}. \tag{10}$$

From the differentials of these equations with respect to ρ and Ψ, it is possible to obtain the fractional errors in normalized resistance and reactance that result from the errors in the determination of VSWR and nodal position. The fractional errors in resistance and reactance due to small incremental errors $\delta\rho$ and $\delta\Psi$ can be expressed as follows:

$$\frac{\delta r}{r} = \frac{1 - \frac{(\rho^2+1)}{2}(1-\cos 2\Psi)}{1 + \frac{(\rho^2-1)}{2}(1+\cos 2\Psi)}\cdot\frac{\delta\rho}{\rho} \tag{11}$$

$$\frac{\delta r}{r} = \frac{(\rho^2-1)\sin 2\Psi}{1 + \frac{(\rho^2-1)}{2}(1+\cos 2\Psi)}\cdot\delta\Psi \tag{12}$$

$$\frac{\delta x}{x} = \frac{\frac{2\rho^2}{(\rho^2-1)}}{1 + \frac{(\rho^2-1)}{2}(1+\cos 2\Psi)}\cdot\frac{\delta\rho}{\rho} \tag{13}$$

$$\frac{\delta x}{x} = \frac{-2\left[1 - \frac{(\rho^2+1)}{2}(1+\cos 2\Psi)\right]}{(\sin 2\Psi)\left[1 + \frac{(\rho^2-1)}{2}(1+\cos 2\Psi)\right]}\cdot\delta\Psi. \tag{14}$$

These equations are plotted in Figs. 5, 6, 7, 8(a), and 8(b). The graphs present the absolute values of the factors by which any fractional errors in the determination of VSWR, and any errors in nodal position, should be multiplied in order to obtain the resulting fractional errors in the load resistance and reactance. The abscissas Ψ of the curves are distances in electrical degrees

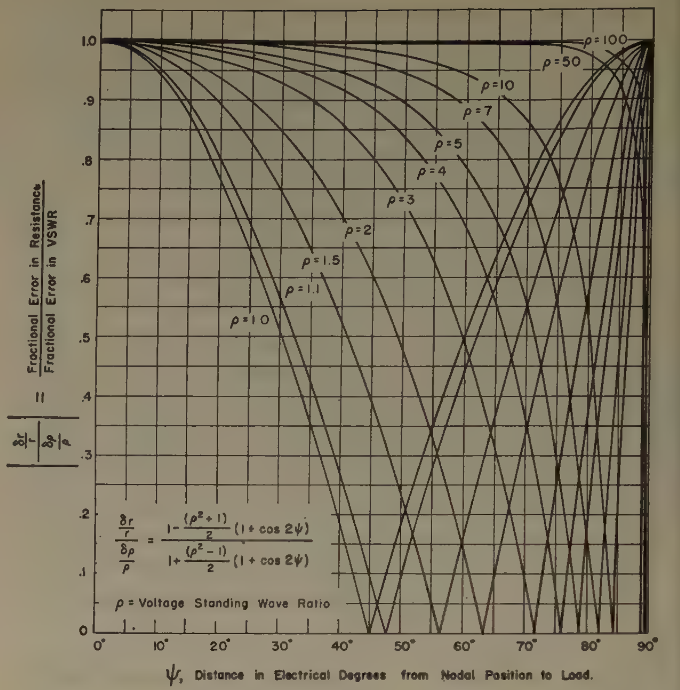


Fig. 5—Curves for determining errors in resistance caused by small errors in measurement of VSWR.

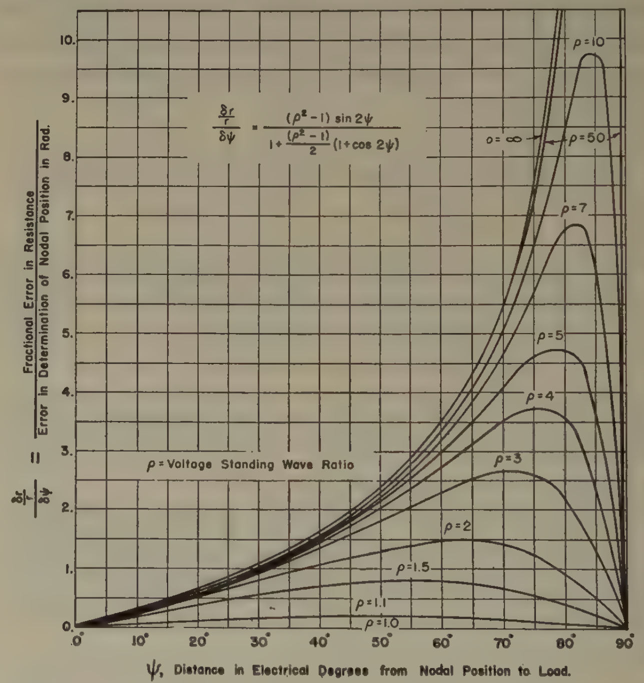


Fig. 6—Curves for determining errors in resistance caused by small errors in measurement of nodal position.

from the load terminals to the nearest voltage minimum. If Ψ is between 90° and 180°, the errors are the same as for an electrical distance of $180° - \Psi$ since the curves are symmetrical about $\Psi = 90°$.

A Smith Chart[15] is useful in understanding the shape of the curves of Figs. 5, 6, 7, 8(a), and 8(b). For example, in Fig. 5 the error in resistance is found to be zero for certain combinations of values of VSWR and nodal

[15] P. H. Smith, "Transmission-line calculator," *Electronics*, vols. 12 and 17; January, 1939, and January, 1944.

position. An inspection of the portion of the Smith Chart which includes these values of VSWR and nodal position shows that the resistance is practically constant for small changes in VSWR. Similarly, in Figs. 8(a) and 8(b), the error in reactance is seen to be zero for certain values of VSWR and nodal position. For these values, the Smith Chart shows the reactance to be practically constant for small changes in nodal position. Fig. 8(b) is a supplement to Fig. 8(a) and presents the ratio $\delta x/\delta\Psi$ for values of Ψ near 0° and 90°. Fig. 8(b) is desirable since the ratio $(\delta x/x)/\delta\Psi$ in Fig. (8a) becomes too large for graphical representation with the selected scale as Ψ approaches 0° or 90°.

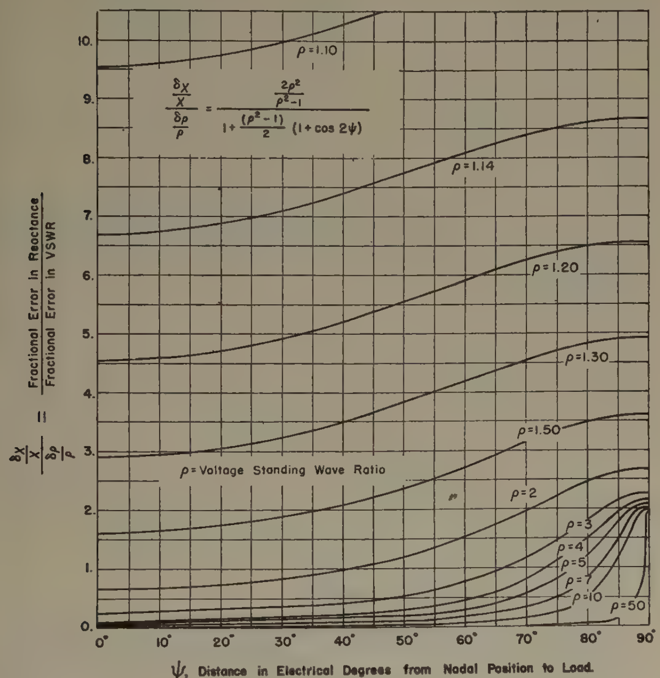
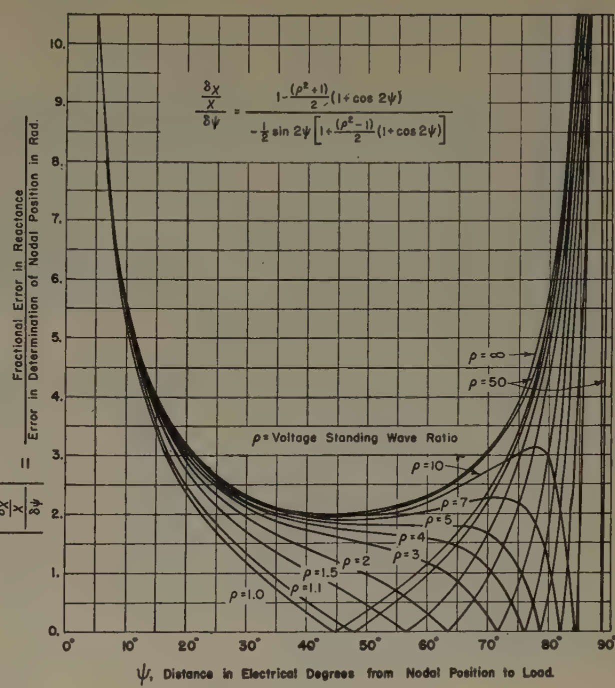
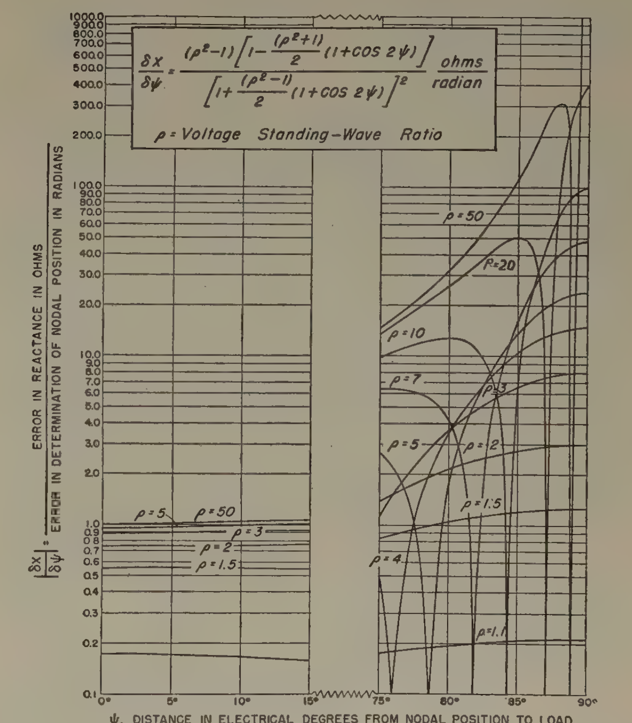


Fig. 7—Curves for determining errors in reactance caused by small errors in measurement of VSWR.



(a)

Fig. 8(a)—Curves for determining errors in reactance caused by small errors in measurement of nodal position.



(b)

Fig. 8 (b)—Supplementary curves for determining errors in reactance caused by small errors in measurement of nodal position for values of Ψ near 0° or 90°.

Conclusion

A study of the accuracy of impedance measurements made by determining the voltage-standing-wave ratio and nodal position in a slotted transmission line indicates that all errors except those introduced by structural imperfections can be corrected or evaluated by known techniques. It is possible to obtain the maximum magnitude of this last type of error by experimentally determining the maximum error in relative-voltage measurement for each standing-wave machine and frequency band. The maximum fractional error in the voltage-standing-wave ratio is twice this constant, and the maximum error in nodal position can then be obtained from a graph, provided the proper equal-response voltages are used.

The effect of these and any other errors in VSWR and nodal position on the accuracy of the calculated terminal resistance and reactance can also be obtained from graphs. Therefore, although the errors in impedance caused by imperfections in the standing-wave machine cannot be easily corrected, their maximum values can be obtained from an experimentally determined constant and a few equations or graphs.

Appendix

A. Maximum Error in Locating Nodal Position

In Fig. 2, Ψ and θ are respectively the distances in electrical degrees from the load terminals to the nodal and probe positions. θ_1 and θ_2 are the probe locations of two equal response voltages. The maximum error $\delta\Psi$ in determining the nodal position is half the sum of the maximum errors, $\delta\theta_1$ and $\delta\theta_2$, in the location of the equal response voltages. The limits of the measured nodal position may be expressed as follows:

$$\Psi \pm \delta\Psi = \frac{(\theta_1 \pm \delta\theta_1) + (\theta_2 \pm \delta\theta_2)}{2} \tag{15}$$

or

$$\Psi \pm \delta\Psi = \frac{\theta_1 + \theta_2}{2} \pm \frac{\delta\theta_1 + \delta\theta_2}{2}. \tag{16}$$

The maximum error in the location of the nodal position is, therefore:

$$\delta\Psi = \pm \frac{\delta\theta_1 + \delta\theta_2}{2}. \tag{17}$$

Expressions for $\delta\theta_1$ and $\delta\theta_2$ can be obtained from the equation for a voltage standing-wave on a lossless transmission line,

$$V = V_{\min}[1 + (\rho^2 - 1)\sin^2(\theta - \Psi)]^{1/2} \tag{18}$$

where $(\theta - \Psi)$ is the distance in electrical degrees from the probe position to the position of a voltage minimum. The maximum error in the location of a selected response voltage is obtained by calculating the distance between positions at which the expressions for V and either $(1+b)V$ or $(1-b)V$ are equal. Because of the symmetry of the voltage distribution curve, the distances $\delta_+\theta_1$ and $\delta_-\theta_2$ of Fig. 2 are equal. Therefore, by substituting $\delta_-\theta_2$ for $\delta_+\theta_1$, $\delta\Psi$ becomes equal to one-half the sum of the absolute values of $\delta_+\theta_2$ and $\delta_-\theta_2$. By equating the expressions for V and $(1-b)V$ at θ_2,

$$(1 - b)[1 - (\rho^2 - 1)\sin^2(\theta_2 - \Psi + \delta_+\theta_2)]^{1/2}V_{\min} = [1 + (\rho^2 - 1)\sin^2(\theta_2 - \Psi)]^{1/2}V_{\min}. \tag{19}$$

Similarly, by equating the expressions for V and $(1+b)V$ at θ_2,

$$(1 + b)[1 + (\rho^2 - 1)\sin^2(\theta_2 - \Psi - \delta_-\theta_2)]^{1/2}V_{\min} = [1 + (\rho^2 - 1)\sin^2(\theta_2 - \Psi)]^{1/2}V_{\min}. \tag{20}$$

From (19),

$$\delta_+\theta_2 = \frac{1}{2}\cos^{-1}\left[\frac{\rho^2+1}{\rho^2-1} - \frac{1 + (\rho^2-1)\sin^2(\theta_2 - \Psi)}{\frac{(\rho^2-1)}{2}(1-b)^2}\right] - (\theta_2 - \Psi). \tag{21}$$

From (20), and the equality of the absolute values of $\delta_+\theta_1$ and $\delta_-\theta_2$,

$$\delta_+\theta_1 = \delta_-\theta_2 = (\theta_2 - \Psi) - \frac{1}{2}\cos^{-1}\left[\frac{\rho^2+1}{\rho^2-1} - \frac{1 + (\rho^2-1)\sin^2(\theta_2 - \Psi)}{\frac{(\rho^2-1)}{2}(1+b)^2}\right] \tag{22}$$

The maximum error, $\delta\Psi$, in the nodal position determined from two equal-response voltages at a distance $(\theta - \Psi)$ electrical degrees from the voltage minimum is equal to half the sum of (21) and (22), or

$$\begin{aligned}\delta\Psi &= \frac{\delta_+\theta_1 + \delta_+\theta_2}{2} \\ &= \frac{1}{4}\cos^{-1}\left[\frac{\rho^2+1}{\rho^2-1} - \frac{1 + (\rho^2-1)\sin^2(\theta - \Psi)}{\frac{(\rho^2-1)}{2}(1-b)^2}\right] \\ &\quad - \frac{1}{4}\cos^{-1}\left[\frac{\rho^2+1}{\rho^2-1} - \frac{1 + (\rho^2-1)\sin^2(\theta - \Psi)}{\frac{(\rho^2-1)}{2}(1+b)^2}\right].\end{aligned} \tag{23}$$

Since $1 + (\rho^2 - 1)\sin^2(\theta - \Psi) = (V/V_{\min})^2$ from (18), $\delta\Psi$ may be expressed as a function of the equal-response voltages used, or

$$\delta\Psi = \frac{1}{4}\cos^{-1}\left[\frac{(\rho^2+1)(1-b)^2 - 2\left[\frac{V}{V_{\min}}\right]^2}{(\rho^2-1)(1-b)^2}\right] - \frac{1}{4}\cos^{-1}\left[\frac{(\rho^2+1)(1+b)^2 - 2\left[\frac{V}{V_{\min}}\right]^2}{(\rho^2-1)(1+b)^2}\right]. \tag{24}$$

B. The Minimum Value of $\delta\Psi$

The value of V for which $\delta\Psi$ is a minimum is obtained by equating to zero the derivative of (24) with respect to V. The value of this equal-response voltage V for which $\delta\Psi$ is a minimum is found to be

$$V = \sqrt{\frac{2\rho^2(1+b^2)}{\rho^2+1}}\, V_{\min}. \tag{25}$$

Since b^2 is very small compared to unity, the expression for V reduces to

$$V = \sqrt{\frac{2\rho^2}{\rho^2+1}}\, V_{\min}. \tag{26}$$

The voltage is of this magnitude at values of $(\theta - \Psi)$ such that

$$\sin(\theta - \Psi) = \sqrt{\frac{\rho^2(1+2b^2) - 1}{\rho^4 - 1}}. \tag{27}$$

If the expression for V in (25) is substituted in (24), the minimum value $\delta\Psi_m$ of the maximum error in determining nodal position is given by

$$\delta\Psi_m = \frac{1}{4}\cos^{-1}\left[1 - \frac{32\rho^2 b^2}{(\rho^2-1)^2(1-b^2)^2}\right]. \tag{28}$$

Alternate Ways in the Analysis of a Feedback Oscillator and its Application*

E. J. POST† AND H. F. PIT†

Summary—**It is a well-known fact that negative feedback has a favorable influence on the phase stability of an amplifier. However, if the negative feedback is applied between the input and output terminals of an amplifier, which is the active part of an oscillating loop, it turns out that the negative feedback may be interpreted either in terms of phase stability of the amplifying section or in terms of phase discriminating properties of the passive frequency determining section of the loop.**

The consequences of this alternate point of view for the design of oscillator networks are discussed.

I. Introduction

An analysis of an oscillator circuit, obtained by writing down the complete set of circuit equations, in principle may give all information available about the behavior of a particular circuit.

Because of the algebraic complexity of the problem, the consequences of changes in circuit elements or the influence of additional elements are not always easily understood.

However, Llewellyn's[1] principles, which make oscillator frequency substantially independent of plate- and filament voltage, may serve as a brilliant example that a complete analysis of the circuit is extremely useful.

From a designer's point of view, some principles allowing a more synthetic approach of the problem may be of use.

Most oscillators may be regarded as a closed loop of an active and a passive four-terminal. In general, the passive four-terminal has a frequency determining function, whereas the active four-terminal is necessary to compensate losses in the frequency determining part.

The problem of generating constant frequencies may be summarized in the conditions which have to be imposed on the active and passive part of the closed loop.

A necessary and sufficient condition for a single closed loop to cause oscillation is given by a Nyquist plot surrounding the appropriate critical point. This criterion, however, gives little information about the frequency which is to be generated.

To study frequency behavior the following necessary, although not sufficient conditions can be used:

1. The net phase-shift ϕ around the oscillating loop must be zero or an integral number of times 2π.

$$\phi = n \times 2\pi, \qquad n = 0, 1, 2, \cdots . \tag{1}$$

2. The net gain g around the closed loop must be one or greater than one:

$$g \geqq 1. \tag{2}$$

In the following text it will be supposed that, moreover, the Nyquist criterion is always satisfied.[2]

Before proceeding to formulate the conditions which have to be imposed on the amplifying and frequency determining part of the closed loop, in order to obtain constant frequencies, we want to call the attention to the fact that the splitting up of a closed loop into an active and passive part is not necessarily unique.

A very instructive example may be illustrated in the following section by means of the network of Meacham's well-known crystal oscillator.[3]

II. A Topological Peculiarity of Meacham's Oscillator[4]

The network shown in Fig. 1(a) is the conventional concept of the Meacham oscillator. In the neighborhood of the generated frequency the tuned transformers will be regarded as ideal. The frequency determining section with in- and output transformer of the amplifier are drawn once more in Fig. 1(b). The dotted line in Fig. 1(b) is of no consequence, in case the connection to the primary of the transformer is suitably chosen no current will flow in it. According to this, the circuit of the Meacham oscillator (Fig. 1(a)) can be rearranged as shown in (Fig. 1(c)). An inspection of Figs. 1(b) and 1(c) shows that the dotted line of 1(b) is a full drawn line in 1(c), connecting a tap of the input transformer to one side of the secondary of the output transformer of the amplifier. Moreover, the resistive branch of the bridge circuit of Fig. 1(a) is included in the amplifier section of Fig. 1(c).

Apparently oscillator networks Figs. 1(a) and 1(c) are equivalent from a point of view of loop transmission.

The only difference between the networks Figs. 1(a) and 1(c) is the sectioning of the loop into amplifying and frequency determining part.

Once the equivalence of networks 1(a) and 1(c) has been established an interesting conclusion may be drawn.

According to Meacham, the frequency stability of oscillator network 1(a) is explained by the extraordinary

* Decimal classification: R139.1×R133. Original manuscript received by the Institute, December 12, 1949; revised manuscript received, July 21, 1950.

† Radio Laboratory of the Netherlands' Postal and Telecommunication Services, 's-Gravenhage, Netherlands.

[1] F. B. Llewellyn, "Constant frequency oscillators," Proc. I.R.E., vol. 19, pp. 2063–2095; December, 1931.

[2] In case of a multiple loop transmission this applies to the intentionally positive feedback loop.

[3] L. A. Meacham, "The bridge-stabilized oscillator," Proc. I.R.E., vol. 26, pp. 1278–1295; October, 1938.

[4] The point of view developed in this section has been suggested by the comments of G. H. Bast and W. H. van Zoest.

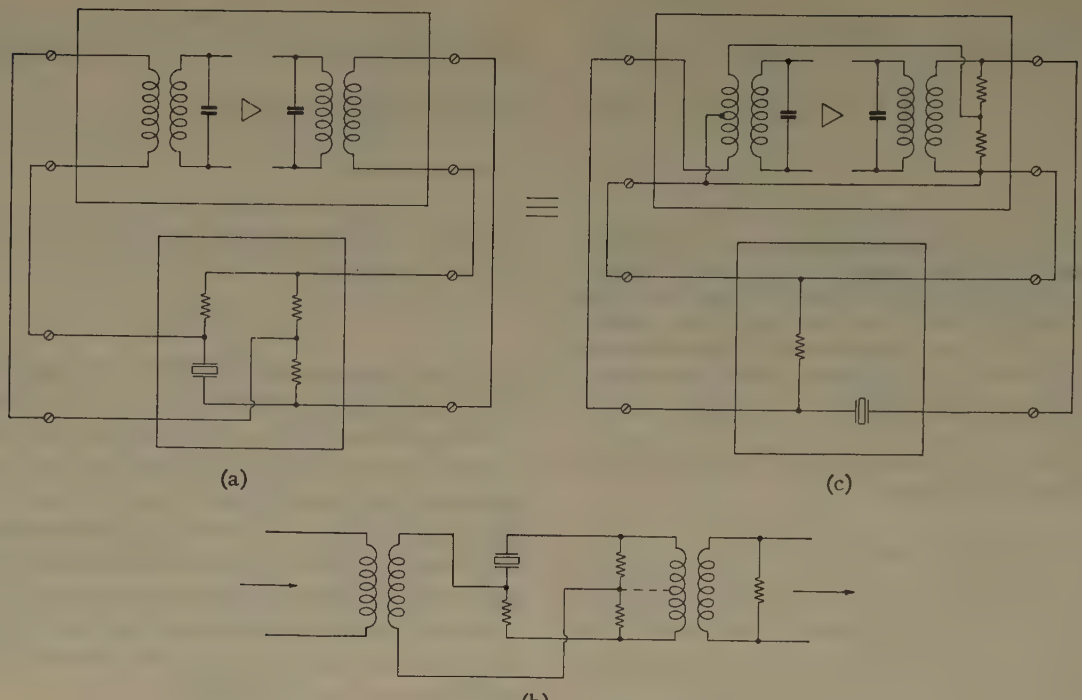


Fig. 1—Network equivalence for Meacham oscillator circuit.

phase discriminating features of the bridge circuit in the frequency determining four-terminal, however, obtained at the expense of additional attenuation.

The frequency determining four-terminal of network 1(c) has a comparatively moderate attenuation, and exhibits no extraordinary phase-discriminating qualities. The amplifier part, on the other hand, has a considerable amount of negative feedback, reducing its excess of gain and resulting in an improved phase stability of the active part of the loop circuit.

With regard to general relations between phase and attenuation, the following rule is suggested:

An additional attenuation associated with a corresponding improvement in the phase discriminating properties of the frequency determining four-terminal may as well be interpreted as an inversed feedback between the output and input terminals of the amplifier.

The rule enunciated above allows one to approach the problem in more than one way, and can be used as a suitable tool in oscillator network synthesis.

In the same way as the bridge circuit can be interpreted in terms of voltage feedback, the bridge-T circuit may be interpreted in terms of current feedback. As discussed by Shepherd and Wise[5] the phase discriminating properties of the Hartley (or Colpitts) oscillator can be improved by the use of a bridged-T section. Figs. 2(a) and 2(b) show that a frequency determining section of a Hartley oscillator connected to an amplifier with current feedback is equivalent to a bridged-T section combined with an amplifier of which the current feedback has been removed.

[5] W. G. Shepherd and R. O. Wise, "Variable-frequency bridge-type frequency-stabilized oscillators," Proc. I.R.E., vol. 31, pp. 256–269; June, 1943.

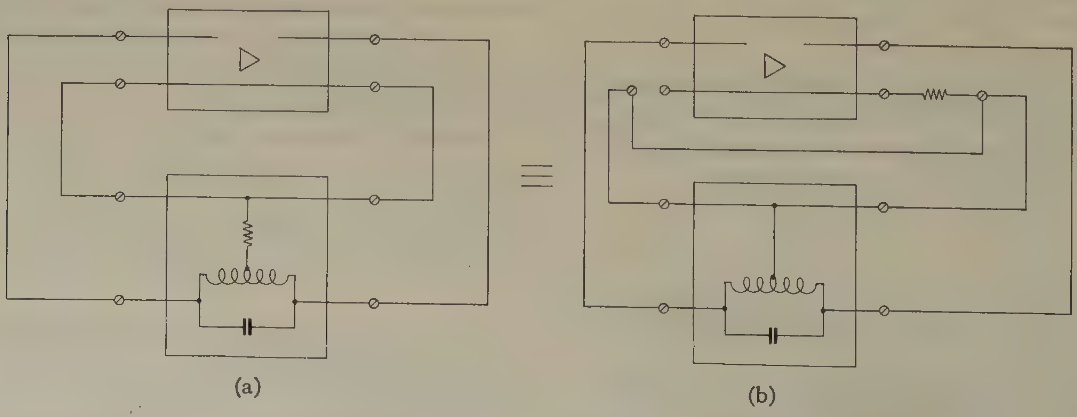


Fig. 2—Network equivalence for Hartley oscillator circuit.

III. Phase Properties of Active and Passive Sections

Having made a definite separation between active and passive section of the closed loop, the attention may be directed to a more detailed study of the phase-frequency characteristics of amplifier $\phi_1(\omega)$ and frequency determining four-terminal $\phi_2(\omega)$ (see Fig. 3). In the neighborhood of the frequency to be generated, ϕ_1 and ϕ_2 may be developed around their points of zero phase[6] ω_1 and ω_2

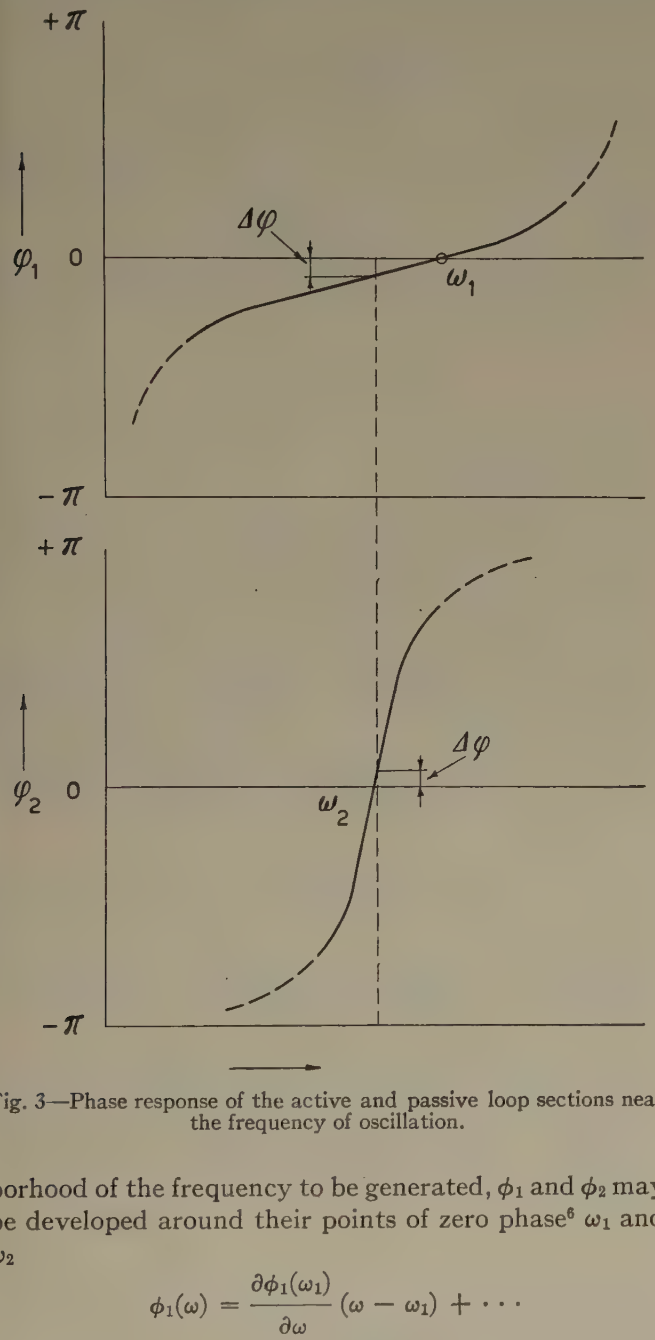


Fig. 3—Phase response of the active and passive loop sections near the frequency of oscillation.

$$\phi_1(\omega) = \frac{\partial\phi_1(\omega_1)}{\partial\omega}(\omega - \omega_1) + \cdots$$

$$\phi_2(\omega) = \frac{\partial\phi_2(\omega_2)}{\partial\omega}(\omega - \omega_2) + \cdots$$

adding and making use of condition 1 (page 169)

$$\phi_1(\omega) + \phi_2(\omega) = 0.$$

[6] Here phase shift eventually with reference to 180°.

Hence

$$\omega = \omega_1 \frac{1}{1+S} + \omega_2 \frac{1}{1 + \frac{1}{S}} \tag{3}$$

if

$$S = \frac{\dfrac{\partial\phi_2(\omega_2)}{\delta\omega}}{\dfrac{\partial\phi_1(\omega_1)}{\delta\omega}}. \tag{4}$$

The natural condition to make the influence of the frequency determining part predominating is:

$$S \gg 1. \tag{5}$$

The magnitude of S, defined in (4), has to be regarded as a criterion to what extent the frequency generated in the loop can be made independent of the amplifier section. However, an excessive high figure for S only is by no means a guarantee for frequency constancy.

It is worth while to point out that the definition of S satisfies the principle of equivalence enunciated in section II. A moderate slope $\partial\phi_2(\omega_2)/\partial\omega$ combined with a very small $\partial\phi_1(\omega_1)/\partial\omega$ of the active section yields a same figure for S as a very large slope $\partial\phi_2(\omega_2)/\partial\omega$ combined with a moderate slope of $\partial\phi_1(\omega_1)/\partial\omega$.

The relations (3) and (5) may serve as a starting point to formulate the prevailing properties which have to be satisfied by the active and passive four-terminal sections of the loop for generation of constant frequencies.

Amplifier A

1. The slope of phase versus frequency must be small and constant in the neighborhood of the generated frequency ω.
2. The frequency of zero phase ω_1 is fixed near ω.

The importance of negative feedback is obvious as a means to diminish the slope of phase versus frequency.[7]

It is believed that, in the long run, the phase stability of capacity resistance coupled amplifiers compare favorably to tuned amplifiers.

Frequency Determining Circuit B

1. The point of zero phase ω_2 must be fixed.
2. The slope of phase versus frequency must be large and constant.

Fluctuating stray capacities across input and output terminals of the frequency determining circuit must have little influence on ω_2. In many cases low impedance filter sections are favorable.[8,9]

[7] F. E. Terman, "Radio Engineer's Handbook," p. 401, Fig. 38, McGraw-Hill Book Co., Inc., New York, N. Y.; 1943.

[8] J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability," Proc. I.R.E., vol. 36, pp. 356–359; March, 1948.

[9] G. F. Lampkin, "An improvement in constant-frequency oscillators," Proc. I.R.E., vol. 27, pp. 199–202; March, 1939

Requirement B_2 involves the use of nearly purely reactive elements, hence high Q. In case of electromechanical elements like crystals, the accessory reactive elements of the four terminals must be chosen with care so that the favorable properties with regard to requirement B_1 are not impaired. For crystal oscillators this is an argument to use series resonance, using the electromechanical branch of the crystal only.

It must be kept in mind that condition B_1 has a priority over condition B_2. An extremely high Q is useless, unless condition B_1 is satisfied. From a designer's point of view it is very instructive to study existing oscillator circuits by splitting up in passive and active sections and analyzing the subsequent parts in order to learn to what extent the conditions mentioned above are satisfied.

IV. Oscillator Circuits

Any properly matched combination of amplifier stages and frequency determining four terminals, possibly on low impedance basis, allowing a net phase shift of zero or $n \times 2\pi$ radians and a loop gain adjustable to unity, may give a suitable, linearly operating, oscillator.

An amplifier section operating in a linear part of its characteristic is an essential condition to guarantee phase stability, hence frequency stability of the complete loop.[7]

An adjustment of the loop gain of a purely linear oscillator to unity exactly is well beyond the aims which can be realized physically, unless an automatically regulating device is included.

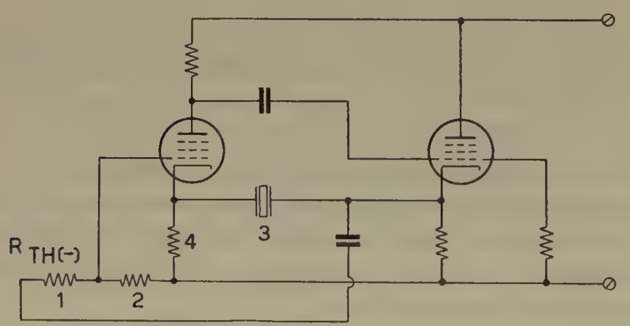


Fig. 4

The oscillator circuit shown in Fig. 4 may serve as an example for the principles of design developed in the foregoing sections.[10] A low impedance output stage of the amplifier has been used to supply the input voltage of the bridge. The positive feedback loop is connected to the cathode, the negative feedback loop to the grid of the differential input stage of the amplifier. The two tubes are coupled by a conventional capacity resistance network to close the loop.

The amplitude controlling device is usually included in the negative feedback loop by means of a thermal resistance. The properties of this system of amplitude control are not identical to those of the conventional Meacham bridge, because of the fact that one of the bridge elements is a dynamic impedance. The influence of this impedance is such that a completely compensating control of the amplitude can be obtained in spite of a very moderate gain of the amplifier.

Moreover, the bridge is not seriously affected if crystals with considerable difference in loss resistance are inserted, in view of the fact that the dynamic cathode impedance of the differential input-stage automatically readjusts the bridge equilibrium.

An analysis of the amplitude controlling device is somewhat lengthy, but straightforward (see Appendix).

V. Conclusion

The alternate points of view discussed in Sections II and III of this paper have been illustrated by methods of "network geometry."[11]

The principal aim of its application has been to emphasize the prevailing points which make oscillator frequency substantially independent of ambient influences on the amplifying section, e.g., variations in power supply, and aging of tubes and circuit elements.

A searching investigation may be necessary to scan the practical boundaries. The difficulties arising from very high figures of S (see (4) and (5)) being used, are analogous and partly identical to the problems encountered in feedback amplifier design and must be attacked accordingly.[12]

Appendix

Analysis of the Amplitude Control

In case one of the bridge elements is the cathode-resistance of an amplifier tube, one has to deal with the effective impedance of the cathode circuit.

For an amplifier stage with a differential input, the input impedance on the cathode is a function of the voltage ratio e_g/e_k, e_g = grid voltage, e_k = cathode voltage. (Fig. 5.) The input impedance Z_i on the cathode is defined as

$$Z_i = \frac{e_k}{I_1} \cdot \tag{6}$$

The circuit equations are

$$e_k = I_1 Z_k - I_2 Z_k \tag{7}$$

$$\mu(e_k - e_g) = I_2(R_i + Z_k + Z_a) - I_1 Z_k. \tag{8}$$

[10] A similar design using two tubes has been advocated by Goldberg and Crosby. The circuit Fig. 4 differs from Goldberg and Crosby's design by the application of negative feedback and by its amplitude controlling features. (H. Goldberg and E. L. Crosby, Jr., *Proc. NEC*, (Chicago), p. 240; November 2-5, 1947.)

[11] As shown by one of the reviewers of the original manuscript, the equivalence is obvious from an analytical point of view as well.

[12] H. W. Bode, "Relations between attenuation and phase in feedback amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, pp. 421-455; July, 1940.

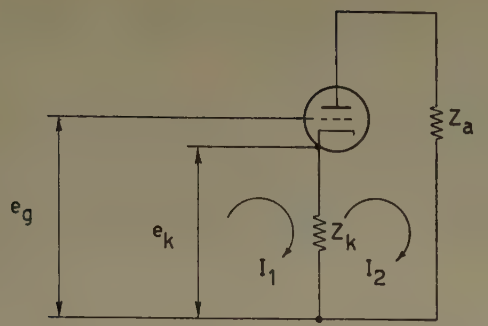


Fig. 5—Amplifier stage with differential input.

Equations (6) and (7) give

$$Z_i = Z_k\left(1 - \frac{I_2}{I_1}\right). \tag{9}$$

Solving (7) and (8) for I_1 and I_2

$$I_1 = -\frac{\mu e_g Z_k - e_k\{R_i + Z_a + (\mu + 1)Z_k\}}{Z_k(R_i + R_a)}$$

$$I_2 = \frac{(\mu + 1)e_k - \mu e_g}{R_i + Z_a}.$$

Substitution in (9), yields an expression for Z_i

$$Z_i = \frac{(R_i + Z_a)Z_k}{R_i + Z_a + Z_k\{1 + \mu(1 - \alpha)\}}. \tag{10}$$

α is written for the voltage ratio e_g/e_k. It is easily verified that $\alpha = 0$ reproduces the input impedance of a grounded grid stage, whereas $\alpha = 1$ (cathode and grid shortcircuited) gives

$$Z_i = \frac{(R_i + Z_a)Z_k}{(R_i + Z_a) + Z_k},$$

expressing that the cathode impedance is shunted by the plate load impedance and internal resistance of the tube in series.

In view of the following application, we will write (10) in the form:

$$R_4 = R_k\frac{R_i + R_a}{R_i + R_a + R_k(1 + \mu(1 - \alpha))}, \tag{10a}$$

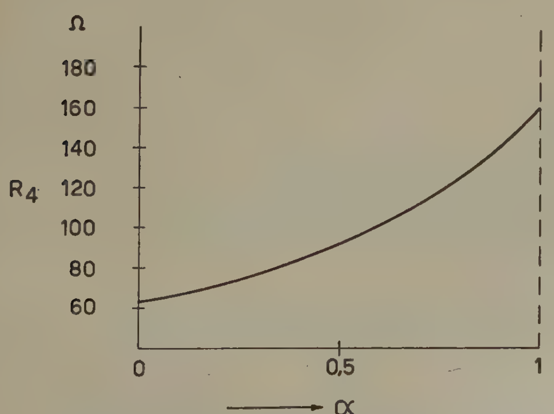


Fig. 6—Cathode input impedance of a differential amplifier stage as a function of voltage ratio α.

in which all the elements are supposed to be purely resistive and R_4 is written for Z_i, being the fourth element of the bridge. Fig. 6 gives a plot for a television pentode of R_4 as a function of the voltage ratio $\alpha = e_g/e_k$, e_g and e_k having the same phase.

The bridge circuit connecting the cathode of the grounded plate stage to the cathode and grid of the differential input stage of oscillator circuit Fig. 4 is shown separately in Fig. 7.

The voltage ratio α may be expressed in the bridge elements $R_{1\ldots4}$; R_3 being the selective series-resistance of the crystal. The resulting expression for $(1-\alpha)$ is (Fig. 7)

$$(1 - \alpha) = \frac{e_k - e_g}{e_k} = \frac{R_1R_4 - R_3R_2}{R_1R_4 + R_2R_4}. \tag{11}$$

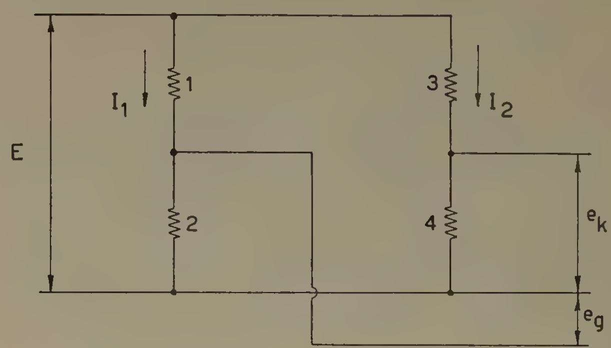


Fig. 7—Bridge circuit connected to the differential amplifier.

Inserting (11) in (10a) gives a linear equation for R_4. Solving for R_4 yields a formula expressing R_4 as a function of the remaining bridge elements R_{1-3} and the parameters of the differential amplifying stage:

$$R_4 = R_k\frac{R_i + R_a + \dfrac{\mu R_2R_3}{R_1 + R_2}}{R_i + R_a + \dfrac{\mu R_1R_k}{R_1 + R_2}} \quad \text{if} \quad R_k \ll (R_i + R_a).$$

Writing S_e for the effective transconductance $S\ R_i/R_i + R_a$, we have the expression:

$$R_4 = R_k\frac{R_1 + R_2 + S_eR_2R_3}{R_1 + R_2 + S_eR_1R_k}. \tag{12}$$

An inspection of equation (12) shows that R_4 approximates

$$\frac{R_2R_3}{R_1} \quad \text{if} \quad S_e \gg \begin{cases} \dfrac{R_1 + R_2}{R_2R_3} \\ \dfrac{R_1 + R_2}{R_1R_k}, \end{cases}$$

or, in other words, the effective resistance R_4 tends to approximate bridge equilibrium.

The practical consequence of this peculiarity is that, within certain limits, crystals with arbitrary differences in loss resistance may be inserted without seriously disturbing the adjustment of the bridge.

It must be emphasized, however, that the phenomenon, being strictly linear, must be looked upon as a device of automatic bridge-readjustment. It has no amplitude controlling features of its own.

A control of the amplitude can be obtained with a nonlinear element only. Let us suppose the bridge element R_1 to be a thermal resistance with negative characteristic. Its nonlinear properties will be explained by the formula

$$R_1 = R_0(1 - \gamma\Delta I_1)$$

now

$$I_1 = \frac{E}{R_1 + R_2} \text{ (see Fig. 7)}$$

and

$$dR_1 = \frac{dR_1}{dE}\,dE = -\frac{\gamma R_0}{R_1 + R_2}\,dE. \tag{13}$$

If the loop is in a steady state of oscillation, the net gain is exactly unity. Therefore the attenuation of the bridge is reciprocal to the gain g of the amplifier. Hence the bridge circuit (Fig. 7) gives an expression for the gain:

$$\frac{e_k - e_g}{E} = \frac{1}{g} = \frac{R_4}{R_3 + R_4} - \frac{R_2}{R_1 + R_2}. \tag{14}$$

Differentiation to the variable elements g, R_1 and R_y gives

$$-\frac{dg}{g^2} = \frac{R_3}{(R_3 + R_4)^2}\,dR_4 + \frac{R_2}{(R_1 + R_2)^2}\,dR_1. \tag{15}$$

According to formula (12) the differential dR_4 can be expressed in the differentials of R_1 and the differential of the gain of the amplifier, because g is proportional to S_e.

$$dR_4 = \frac{\partial R_4}{\partial g}\,dg + \frac{\partial R_4}{\partial R_1}\,dR_1. \tag{16}$$

Inserting (16) and (13) in (15) and solving for dE,

$$dE = -\frac{\dfrac{1}{g^2} + \dfrac{\partial R_y}{\partial g}\dfrac{R_3}{(R_3 + R_4)^2}}{\dfrac{dR_1}{dE}\left[\dfrac{R_3\dfrac{\partial R_4}{\partial R_1}}{(R_3 + R_4)^2} + \dfrac{R_2}{(R_1 + R_2)^2}\right]}\,dg.$$

This expression relates a change in gain to a corresponding change in E, the input voltage of the bridge. A compensated control of the amplitude is possible if

$$\frac{\partial R_4}{\partial g} = -\frac{(R_3 + R_4)^2}{g^2 R_3}. \tag{17}$$

An inspection of formula (17) shows that $\partial R_4/\partial g < 0$ if $R_k R_1 \gg R_2 R_3$. The last inequality being a necessary condition to assure an excess of positive feedback.

For the computation of the differential quotient $\partial R_4/\partial g$ it has to be remembered that the gain of the amplifier depends on S_e. There are other factors affecting the gain; the influence, however, of the second (cathode follower) tube is comparatively small because of its individual inversed feedback.

A similar conclusion is obtained if R_2 is taken as a thermal resistance with positive characteristics.

Hence, the combined efforts of pseudo-linear element and dynamic impedance as elements in the bridge create the possibility of a compensated control of the amplitude for a finite and very moderate gain of the amplifier.

In general, the source supplying the input voltage of the bridge has a finite internal impedance. Therefore, the use of a thermistor (negative characteristic) being a constant voltage device on its own, assists the stabilizing properties of the control action.

Correction

D. D. King, author of the paper, "Two standard field-strength meters for very-high frequencies," which appeared on pages 1048–1051 of the September, 1950, issue of the Proceedings of the I.R.E., has brought the following error to the attention of the editors.

In the caption for Fig. 7 on page 1051, "1 volt per wavelength ($E\lambda_r = 1$)," should read "1 volt per resonant wavelength ($E\lambda_r = 1$)."

Spectrum Analysis of Transient Response Curves*

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Summary—**The paper describes a method of computation of amplitude and phase response of a network from its measured transient response. Tables and nomographs have been included to facilitate the numerical evluation.**

Introduction

SINCE TRANSIENT tests are rapidly becoming standard engineering tools, the importance of spectrum analysis of measured response curves is steadily increasing.

It is well known that such an analysis can be made in cases where the mathematical expression of the respective curve is given by evaluating the Fourier integral. In cases where only the graph of the response curve is known, this leads to a lengthy and laborious graphical integration and is hardly practical.

Another method will be briefly described here which was presented by Bedford and Fredendall.[1] The principle of this method, called by its authors "square-wave analysis," is as follows:

The given curve is approximated by the sum of several step-functions which are regularly spaced in time (Fig. 1). The complex Fourier spectrum of each step function is computed. The sum of the individual spectra

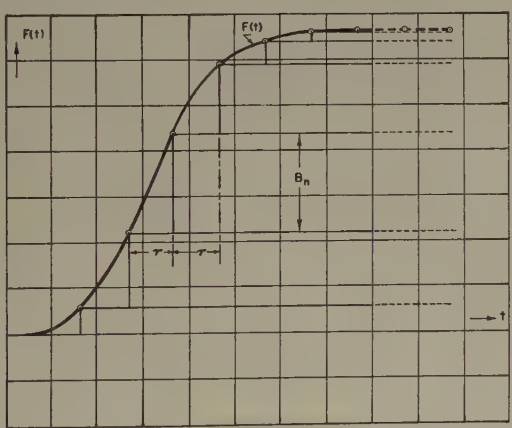

Fig. 1—Stair-step approximation of a given curve.

is the spectrum of the sum of step functions. Since this sum is assumed to be a good approximation of the given curve, it is reasonable to assume that its spectrum is a good approximation of the spectrum of the original curve. The corresponding mathematical operations are given below: If $F(t)$ is the given curve, $f(t)$ a unit step function and τ the spacing between adjacent steps:

$$F(t) \approx \sum_{n=0}^{\infty} B_n f(t - n\tau) \tag{1}$$

where B_n are the amplitudes of the individual step functions.

The Fourier spectrum of a single step function is:

$$\phi(\omega) = \int_{-\infty}^{+\infty} B_n f(t - n\tau) e^{-j\omega t} dt = B_n \frac{e^{-jn\omega\tau}}{j\omega} \tag{2}$$

and the spectrum of the function $F(t)$:

$$\Phi \approx \frac{1}{j\omega} \sum_{n=0}^{\infty} B_n e^{-jn\omega\tau}. \tag{3}$$

The equation for the (complex) transfer function $H(\omega)$, if we assume an ideal unit function at the input terminals, is, therefore:

$$H(\omega) \approx \sum_{n=0}^{\infty} B_n e^{-jn\omega\tau}. \tag{4}$$

Method Using (sin x/x)-Functions

This method which has proved very useful in a great number of cases is based on the following fact: Almost all transient response curves have one property in common; namely, their spectrum does not contain components above a certain limit, say f_c, to any appreciable amount. This is due to the inherent properties of the system under test and of the test equipment itself. In cases where an oscilloscope is used, the scope amplifier might be the limiting item; in other cases, different parts of the test setup might cause the frequency limitation. A function of this type, namely, a function the frequency spectrum of which has an upper limit f_c, can be synthesized exactly by a sum of functions of the type sin x/x. This was proved by Shannon in a recent pub-

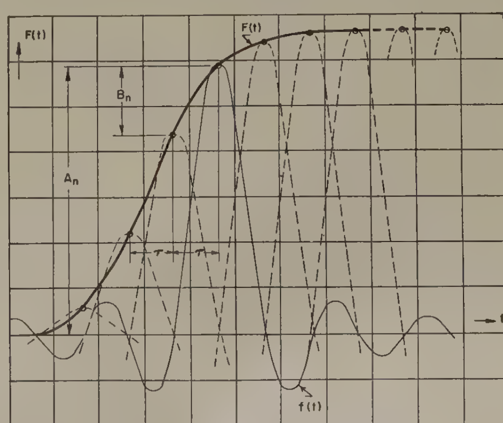

Fig. 2—Synthesis of a given curve by sin x/x functions.

* Decimal classification: R143×537.7. Original manuscript received by the Institute, March 15, 1950; revised manuscript received, August 22, 1950. Presented, 1949 National IRE Convention, New York, N. Y., March 8, 1949.

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[1] A. V. Bedford and G. L. Fredendall, "Analysis, synthesis, and evaluation of the transient response of television apparatus," Proc. I.R.E., vol. 30, pp. 440–458; October, 1942.

lication.[2] The conditions for the synthesis will be explained with the help of Fig. 2: The measured curve is "sampled" at regular intervals (τ) which must be smaller than or equal to ($1/2f_c$). At each sample point, a curve of the described type is placed with an amplitude equal to the amplitude of the corresponding sample point. In Fig. 2 the sin x/x curves are only partly drawn in order to improve the clarity of the figure. The next step is to find the complex Fourier spectrum of each individual

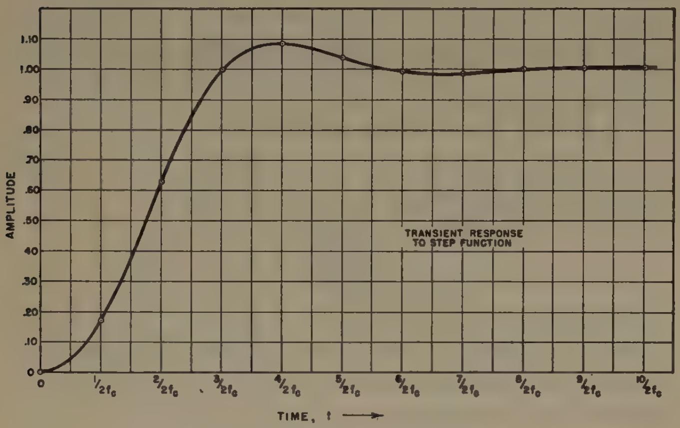


Fig. 3—Transient response curve.

sin x/x function. Then, the sum of these spectra is the Fourier spectrum of the measured curve. Provided that the above assumption holds, that is, that no frequency components above fc are present, this would be the correct answer, not an approximation. The validity of this assumption will be discussed later.

The mathematical derivations for the method described above, follow: Let the given time function be again $F(t)$ and τ the time spacing between adjacent sampling points, then:

$$F(t) = \sum_{n=0}^{\infty} A_n \frac{\sin f_c(t - n\tau)2\pi}{f_c(t - n\tau)2\pi}; \quad \tau \leqq \frac{1}{2f_c} \tag{5}$$

where A_n are the amplitudes of the sampling points, and f_c is the highest component in the frequency spectrum of $F(t)$. It can be shown that the spectrum of the nth term of the sum of (5) is

$$\phi(\omega) = \int_{-\infty}^{+\infty} A_n \frac{\sin 2\pi f_c(t - n\tau)}{2\pi f_c(t - n\tau)} e^{-j\omega t} dt$$

$$= A_n \frac{e^{-jn\tau\omega}}{2f_c}. \tag{6}$$

Hence, the Fourier spectrum of the given curve $F(t)$ is:

$$\Phi(\omega) = \sum_{n=0}^{\infty} A_n \frac{e^{-jn\omega\tau}}{2f_c}; \qquad \tau = \frac{1}{2f_c}. \tag{7}$$

[2] C. Shannon, "Communications in the presence of noise," Proc. I.R.E., vol. 37, pp. 10–22; January, 1949.

This equation relates the spectrum to the measured points $A_0, A_1, A_2 \cdots$ of the transient response curve and is, therefore, the answer to our problem. In many cases, however, it is of advantage to express the spectra not by the sample amplitudes A, but by the first differences or increments of A. If we define $B_n = A_n - A_{n-1}$, and neglect constant time delay, then (7) becomes (8), as derived in the mathematical Appendix.

$$\Phi(\omega) = \frac{1}{j4f_c \sin \frac{\pi}{2} f/f_c} \sum_{n=0}^{\infty} B_n e^{-jn\pi f/f_c}. \tag{8}$$

In a case where the measured curve is the response to an ideal step function the complex transfer function of the system under test will be described by (9). The transfer function is, of course, the complex notation of amplitude and phase response of a network.

$$H(\omega) = \frac{\pi/2 \cdot f/f_c}{\sin(\pi/2 \cdot f/f_c)} \sum_{n=0}^{\infty} B_n e^{-jn\pi f/f_c}. \tag{9}$$

Our assumption, on which the derived formulas (8) and (9) are based, was that the spectrum of the response curve does not contain frequencies above a certain limit f_c. Since this assumption will never be fulfilled with full mathematical rigor for a practical network, the practical value of the method might be questioned. It can, however, be shown that the amount of the error, which will be largest near the nominal cutoff frequency f_c, will be in general smaller than the amplitude response at f_c, provided that the response does not rise again above its value at f_c for frequencies $f/f_c > 1$. An example, in which the condition for a frequency cutoff has been fulfilled only approximately, follows here.

A certain transient response curve (Fig. 3) was assumed of which the accurate mathematical expression is given by (10)

$$F(t) = 1 - e^{-2\pi f_0 t} - \frac{2}{\sqrt{3}} e^{-\pi f_0 t} \cdot \sin(\sqrt{3}\,\pi f_0 t) \tag{10}$$

and analyzed by three different methods: First, by the evaluation of the Fourier-integral which, of course, provides the exact answer;[3] second, by the method just described and, third, by the "square-wave method" mentioned earlier. The time spacing for the sampling was chosen to be $1/5f_0$. The results are shown in Fig. 4. The solid curve shows the correct solution, the crosses the results of the proposed method and the circles the results of the "square-wave method." The time delay-response is omitted for reasons of space.

In spite of the fact that our assumption was not completely fulfilled (we have at f_c only 24 db instead

[3] The solution is:

$$H(\omega) = \frac{1}{-(f/f_0 - e^{+j\pi/6})(f/f_0 + e^{-j\pi/6})(1 + jf/f_0)}.$$

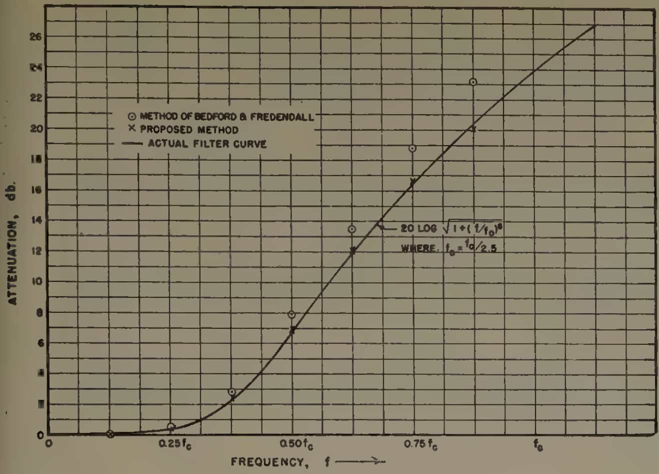


Fig. 4—Amplitude response corresponding to transient response of Fig. 3.

of an infinitely large attenuation), the results obtained are of considerable accuracy. This can also be seen from Fig. 5, which shows the relative error of the amplitude responses.

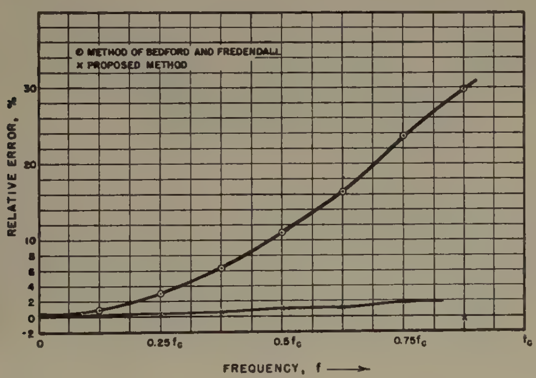


Fig. 5—Relative error of amplitude response.

It is believed that for the same accuracy the proposed method allows a considerable decrease of the number of sampling points compared with previous methods and, therefore, constitutes a time saving in the numerical evaluation.

It might be mentioned that in a case where the natural cutoff frequency f_c is far above the frequency of interest, the cutoff can be lowered artificially by the use of a low-pass filter. One can thus avoid an unnecessarily large number of sampling points.

The Numerical Evaluation

In this paragraph it will be shown how the numerical evaluation of the formula for the transfer function can be made. Since the following simple relationship exists between the transfer function $H(\omega)$ and the spectrum (assuming an ideal step function input),

$$H(\omega) = \frac{\Phi(\omega)}{E/j\omega} \tag{11}$$

where

$H(\omega)$ = complex transfer function
$\Phi(\omega)$ = complex Fourier spectrum
$E/j\omega$ = complex spectrum of an ideal input step function of amplitude E

it is not necessary to discuss the numerical evaluation of $\Phi(\omega)$ as well.

For a certain frequency f_0 for which the calculation of phase and amplitude of a network is desired, the expression for $H(\omega)$ takes the form:

$$H(\omega_0) = \frac{\pi/2\cdot f_0}{f_c \sin\frac{\pi}{2} f_0/f_c} \cdot \sum_{n=0}^{\infty} B_n e^{-jn\pi f_0/f_c}. \tag{12}$$

On the right side, all values are known; the B's are the differences of the sample amplitudes, f_0 is the frequency for which the evaluation is being performed, and f_c is the cutoff frequency. The main part of the evaluation is to form the complex sum $\sum B_n e^{-jn\pi f_0/f_c}$. This can be done graphically (as for a similar case proposed by Bedford and Fredendall).[1] Another method is used here; we write:

$$\sum_{n=0}^{\infty} B_n e^{-jn\pi f_0/f_c}$$

$$= \sum_{n=0}^{\infty} B_n \cos(n\pi f_0/f_c) - j\sum B_n \sin(n\pi f_0/f_c) \tag{13}$$

$$\sum B_n e \cdots = \sum_0^{\infty} c_{n_{\cos}} - j\sum_0^{\infty} c_{n_{\sin}}. \tag{14}$$

Tables are computed for different values of n, B_n and f_0/f_c which allow one to find $c_{r_{\sin}}$ and $c_{n_{\cos}}$. If the sums $\sum c_{\sin}$ and $\sum c_{\cos}$ are found and called $C_{\sin}$ and $C_{\cos}$ the nomographs I or II allow one to find the amplitude response of the network being analyzed ($|H(\omega)|$) for f_0. Nomograph III gives the phase for f_0.

It should be mentioned that the phase angle thus obtained is the so-called "principal value," i.e., it might be necessary to add a multiple of π to it.

The constant time delay cannot be computed by this method, since terms pertaining to it were neglected in the course of the derivation.

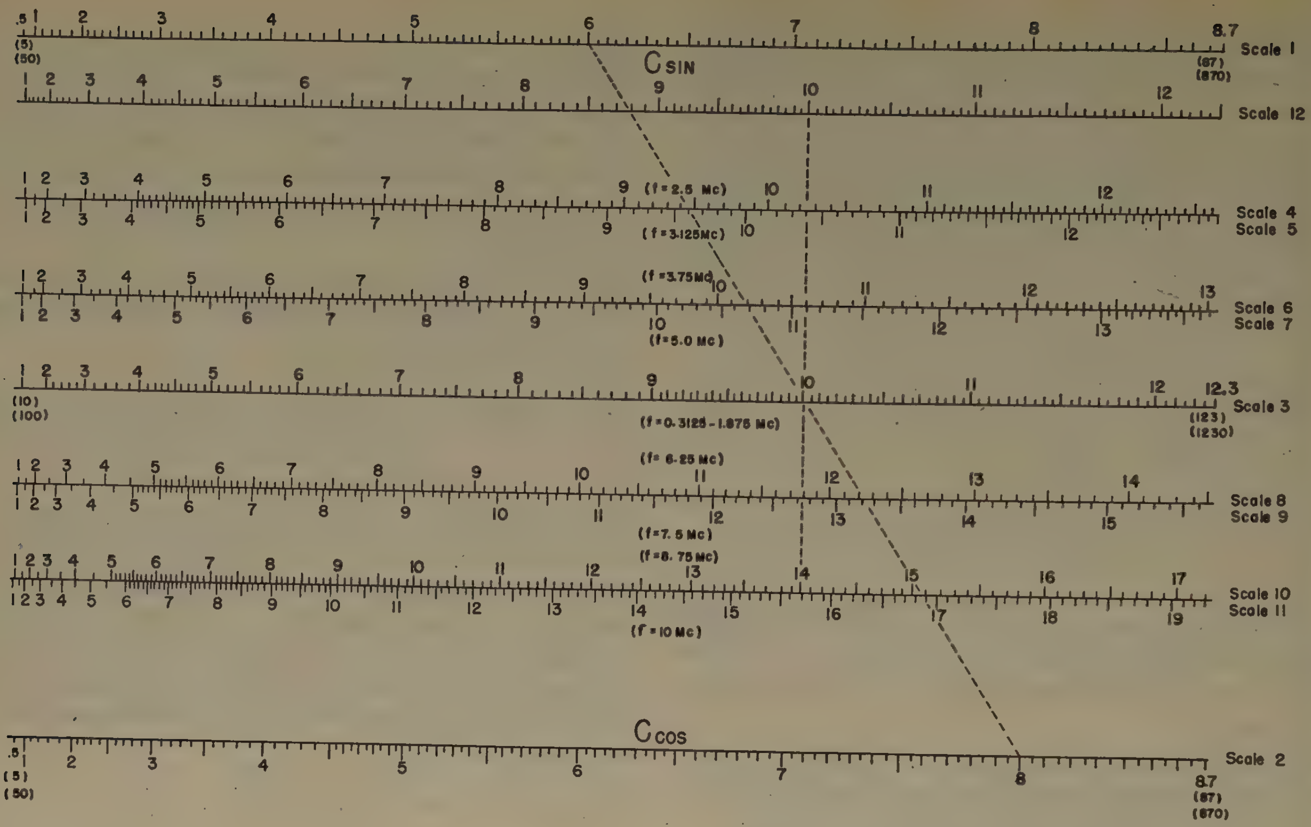
Instructions for the Use of Tables and Nomographs

General

The tables are computed for a sampling point spacing of 0.05 μs; they are valid for any other sampling point spacing τ^* if the frequency values printed on tables and nomographs are multiplied by $0.05/\tau^*\mu$s.

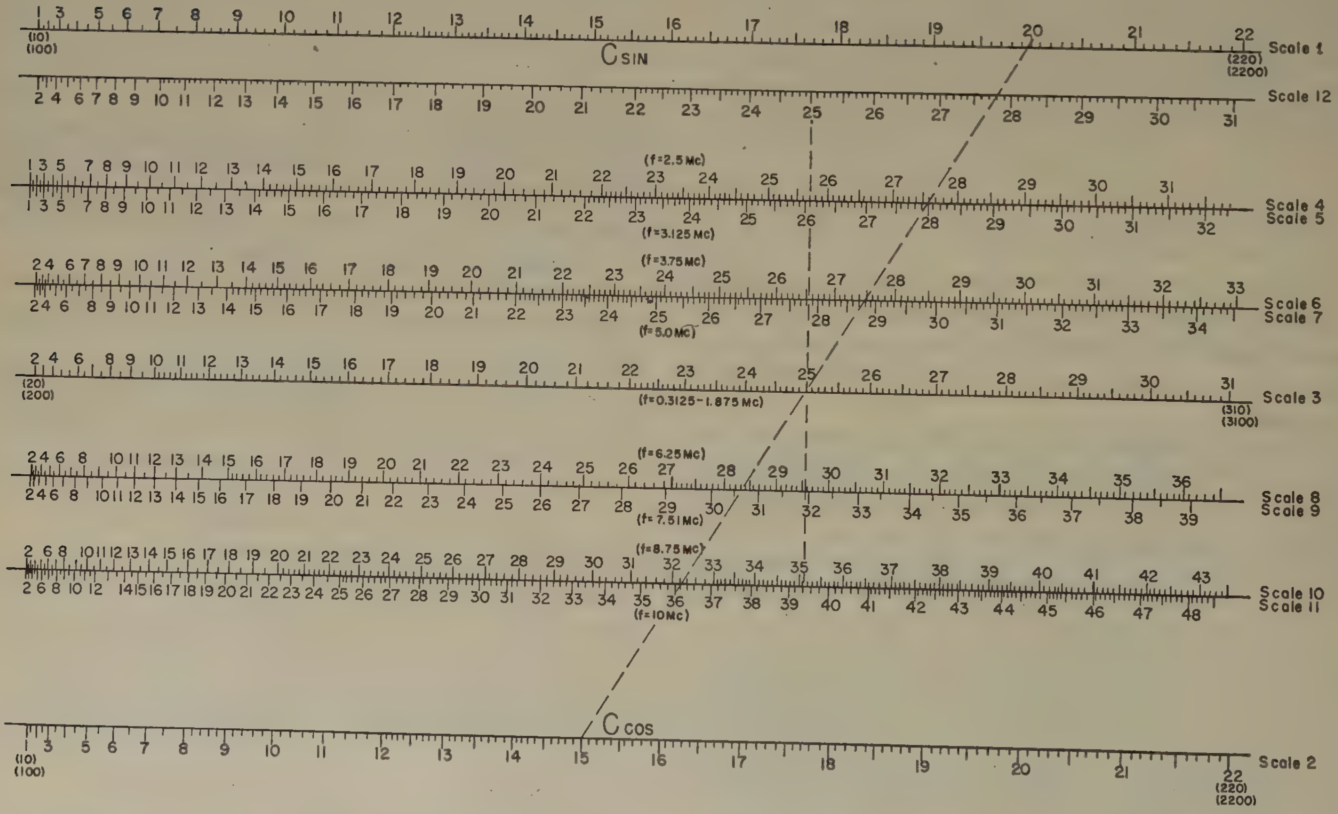
The Sampling

(a) Use a spacing τ smaller than $1/2\, f_c$ where f_c is the cutoff frequency, defined in the first part of this paper.



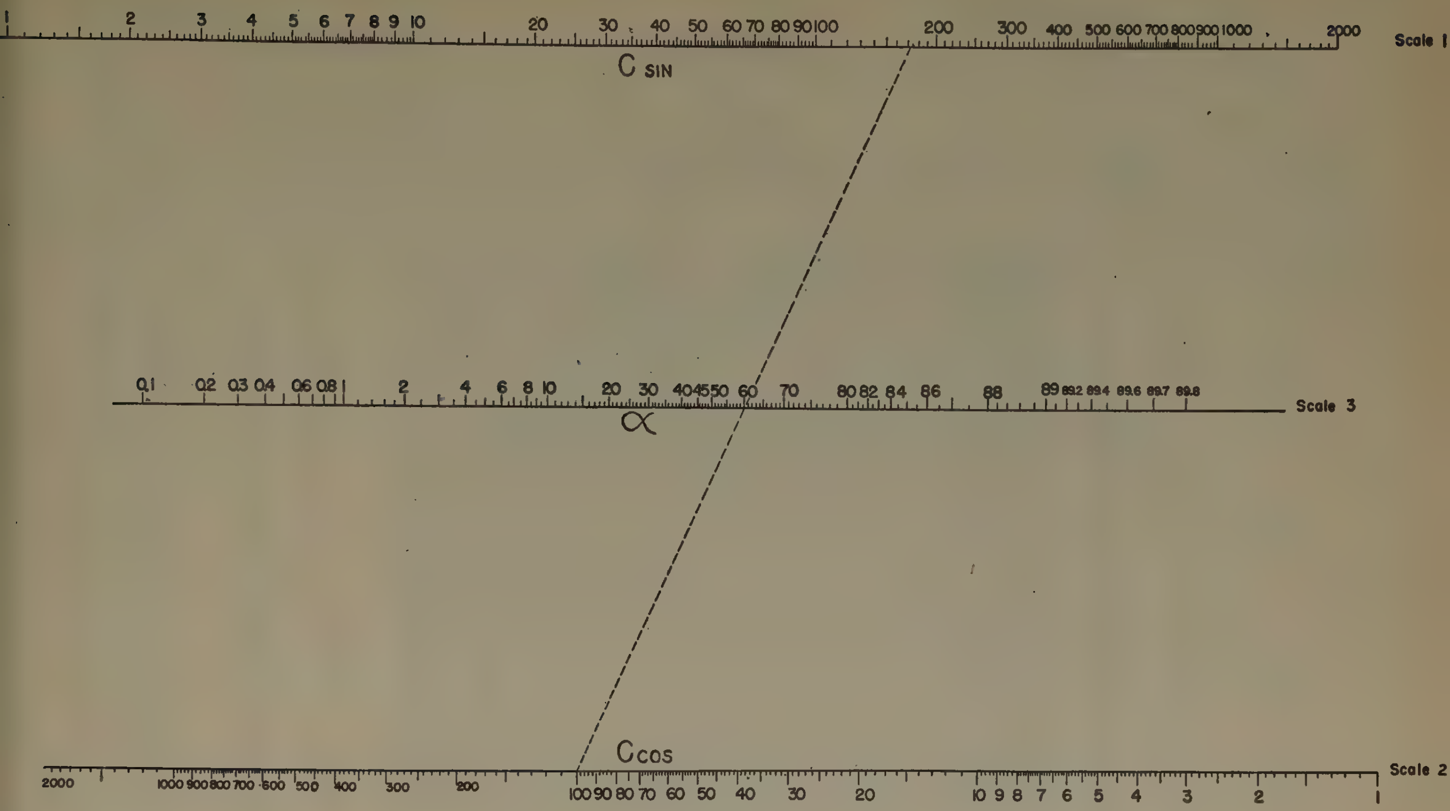
NOTE: TO ESTABLISH THE LINE PERPENDICULAR TO SCALE 3 AND ANY VALUE, CONNECT THIS VALUE ON SCALE 3 WITH THE SAME VALUE ON SCALE 12.

Nomograph I



NOTE: TO ESTABLISH THE LINE PERPENDICULAR TO SCALE 3 AT ANY VALUE, CONNECT THIS VALUE ON SCALE 3 WITH THE SAME VALUE ON SCALE 12.

Nomograph II



Nomograph III

(b) Place the sampling points so that the first point (subscript $n=0$) is not farther to the right than one interval τ past the start of the variation of the curve (see Fig. 6). Sampling points must continue to the point where the curve ceases varying.

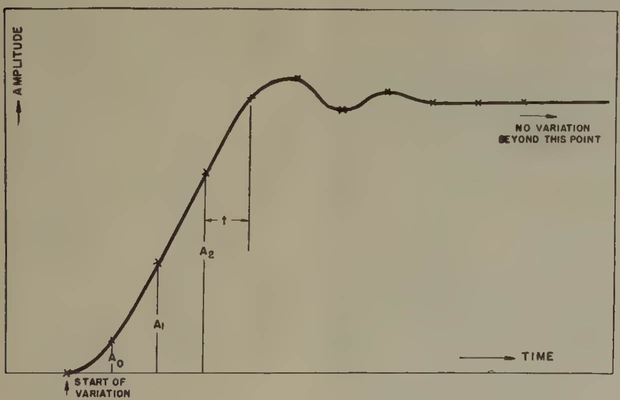


Fig. 6—Sampling of an experimentally determined transient response curve.

(c) Tabulate the amplitudes at the sampling points. Let these be called $A_0, A_1, A_2, A_3, \cdots, A_n, \cdots$.

(d) Form the differences $A_0-0, A_1-A_0, A_2-A_1, \cdots$, and let these be called $B_0, B_1, B_2, \cdots, B_n, \cdots$. (Some values of B might be negative.) If necessary, the values of B should be multiplied by an arbitrary scale factor so that the largest value of B lies between 70 and 99; this assures higher accuracy in the results.[4]

(e) Round off the values of B thus found to the nearest whole number.

The Computation of the Auxiliary Values $C_{\sin}$ and $C_{\cos}$

Select the table (Tables I–XII) for the frequency for which amplitude and phase is to be computed. For calculation of $C_{\sin}$ use the heading marked "for $C_{\sin}$." Enter the argument column (under the heading B_n) with the B_n value; proceed horizontally to that column for which the heading contains the corresponding subscript number N and tabulate the value thus obtained. (*Note:* The value $c_{n_{\sin}}$ which has been found, normally has the same sign as B_n unless the subscript in the heading is preceded by an asterisk, in which case they have opposite sign.) This must be done for all B_n's of which the subscripts are listed in the heading.

It is advisable to perform this conversion not in the sequence of the points (i.e., $B_0, B_1, B_2, \cdots$) but rather in the sequence of the groups indicated by the headings (e.g., for 0.625 Mc and $C_{\sin}$: first for the points with the subscripts 1, 15, 17, 31 then for the next group 2, 14, 18, 30, etc.).

Add all values $c_{n_{\sin}}$ thus found, retaining proper signs. The sum is the auxiliary value $C_{\sin}$.

[4] The scale factor will only appear in the final result for the amplitude; it can be eliminated by dividing the result by the scale factor. This, however, is not necessary if only relative values are of interest.

F=.3125 MC

B_N	N: 1 31 (FOR C_{SIN}) / 15 *17 (FOR C_{COS})	2 30 / 14 *18	3 29 / 13 *19	4 28 / 12 *20	5 27 / 11 *21	6 26 / 10 *22	7 25 / 9 *23	8 24 / 8 *24	9 23 / 7 *25	10 22 / 6 *26	11 21 / 5 *27	12 20 / 4 *28	13 19 / 3 *29	14 18 / 2 *30	15 17 / 1 *31	16 / 0
1	00.10	00.20	00.30	00.38	00.47	00.56	00.63	00.71	00.77	00.83	00.88	00.92	00.96	00.98	01.00	
2	00.20	00.39	00.58	00.77	00.94	01.11	01.27	01.41	01.55	01.66	01.76	01.85	01.91	01.96	01.99	
3	0029	00.59	00.87	01.15	01.41	01.67	01.90	02.12	02.32	02.49	02.65	02.77	02.87	02.94	02.99	
4	00.39	00.78	01.16	01.53	01.89	02.22	02.54	02.83	03.09	03.33	03.53	03.70	03.83	03.92	03.98	
5	00.49	00.98	01.45	01.91	02.36	02.78	03.17	03.54	03.87	04.16	04.41	04.62	04.78	04.90	04.98	
6	00.59	01.17	01.74	02.30	02.83	03.33	03.81	04.24	04.64	04.99	05.29	05.54	05.74	05.88	05.97	
7	00.69	01.37	02.03	02.68	03.30	03.89	04.40	04.95	05.41	05.82	06.17	06.47	06.70	06.86	06.97	
8	00.78	01.56	02.32	03.06	03.77	04.44	05.08	05.66	06.18	06.05	07.06	07.39	07.66	07.85	07.96	
9	00.88	01.76	02.61	03.44	04.24	05.00	05.71	06.36	06.96	07.48	07.94	08.32	08.61	08.83	08.96	
10	00.98	01.95	02.90	03.83	04.71	05.56	06.34	07.07	07.73	08.32	08.82	09.24	09.57	09.81	09.95	
11	01.08	02.15	03.19	04.21	05.19	06.11	06.98	07.78	08.50	09.15	09.70	10.16	10.53	10.79	10.95	
12	01.18	02.34	03.48	04.60	05.66	06.67	07.61	08.49	09.28	09.98	10.58	11.09	11.48	11.77	11.94	
13	01.27	02.54	03.77	04.98	06.13	07.22	08.25	09.19	10.05	10.81	11.46	12.01	12.44	12.75	12.94	
14	01.37	02.73	04.06	05.36	06.60	07.78	08.88	09.90	10.82	11.64	12.35	12.93	13.40	13.73	13.93	
15	01.47	02.93	04.35	05.74	07.07	08.33	09.52	10.61	11.60	12.47	13.23	13.86	14.35	14.71	14.93	
16	01.57	03.12	04.64	06.12	07.54	08.89	10.15	11.31	12.37	13.30	14.11	14.78	15.31	15.69	15.92	
17	01.67	03.32	04.94	06.51	08.01	09.45	10.78	12.02	13.14	14.14	14.99	15.71	16.27	16.67	16.92	
18	01.74	03.51	05.23	06.89	08.49	10.00	11.42	12.73	13.91	14.97	15.87	16.63	17.22	17.65	17.91	
19	01.86	03.71	05.52	07.27	08.96	10.56	12.05	13.43	14.69	15.80	16.76	17.55	18.18	18.64	18.91	
20	01.96	03.90	05.81	07.65	09.43	11.11	12.69	14.14	15.46	16.63	17.64	18.48	19.14	19.62	19.90	
21	02.06	04.10	06.10	08.04	09.90	11.67	13.32	14.85	16.23	17.46	18.52	19.40	20.09	20.60	20.90	
22	02.16	04.29	06.39	08.42	10.37	12.22	13.96	15.56	17.01	18.29	19.40	20.33	21.05	21.58	21.89	
23	02.25	04.49	06.68	08.80	10.84	12.78	14.59	16.26	17.78	19.12	20.28	21.25	22.01	22.56	22.89	
24	02.35	04.68	06.97	09.18	11.31	13.33	15.23	16.97	18.55	19.96	21.17	22.17	22.97	23.54	23.88	
25	02.45	04.88	07.26	09.57	11.79	13.89	15.86	17.68	19.33	20.79	22.05	23.10	23.92	24.52	24.88	
26	02.55	05.07	07.55	09.95	12.26	14.45	16.49	18.38	20.10	21.62	22.93	24.02	24.88	25.50	25.88	
27	02.65	05.27	07.84	10.33	12.73	15.00	17.13	19.09	20.87	22.45	23.81	24.95	25.84	26.48	26.87	
28	02.74	05.46	08.13	10.72	13.20	15.56	17.76	19.80	21.64	23.28	24.69	25.87	26.79	27.46	27.87	
29	02.84	05.66	08.42	11.10	13.67	16.11	18.40	20.51	22.42	24.11	25.58	26.79	27.75	28.44	28.86	
30	02.94	05.85	08.71	11.48	14.14	16.67	19.03	21.21	23.19	24.95	26.46	27.72	28.71	29.42	29.86	
31	03.04	06.05	09.00	11.86	14.61	17.22	19.67	21.92	23.96	25.78	27.34	28.64	29.66	30.40	30.85	
32	03.14	06.24	09.29	12.25	15.08	17.78	20.30	22.63	24.74	26.61	28.22	29.56	30.62	31.39	31.85	
33	03.23	06.44	09.58	12.63	15.56	18.33	20.94	23.33	25.51	27.44	29.10	30.49	31.58	32.37	32.84	
34	03.33	06.63	09.87	13.01	16.03	18.89	21.57	24.04	26.28	28.27	29.98	31.41	32.53	33.35	33.84	
35	03.43	06.83	10.16	13.39	16.50	19.45	22.20	24.75	27.06	29.10	30.87	32.34	33.49	34.33	34.83	
36	03.53	07.02	10.45	13.78	16.97	20.00	22.84	25.46	27.83	29.93	31.75	33.26	34.45	35.31	35.83	
37	03.63	07.22	10.74	14.16	17.44	20.56	23.47	26.16	28.60	30.77	32.63	34.18	35.41	36.29	36.82	
38	03.72	07.41	11.03	14.54	17.91	21.11	24.11	26.87	29.37	31.60	33.51	35.11	36.36	37.27	37.82	
39	03.82	07.61	11.32	14.92	18.38	21.67	24.74	27.58	30.15	32.43	34.39	36.03	37.32	38.25	38.81	
40	03.92	07.80	11.61	15.31	18.86	22.22	25.38	28.28	30.92	33.26	35.28	36.96	38.28	39.23	39.81	
41	04.02	08.00	11.90	15.69	19.33	22.78	26.01	28.99	31.09	34.09	36.16	37.88	39.24	40.21	40.80	
42	04.12	08.19	12.19	16.07	19.80	23.34	26.64	29.70	32.47	34.92	37.04	38.80	40.19	41.19	41.80	
43	04.21	08.39	12.48	16.46	20.27	23.69	27.28	30.41	33.24	35.75	37.92	39.73	41.15	42.17	42.79	
44	04.31	08.58	12.77	16.84	20.74	24.45	27.91	31.11	34.01	36.59	38.80	40.65	42.10	43.16	43.79	
45	04.41	08.78	13.06	17.22	21.21	25.00	28.55	31.82	34.79	37.42	39.68	41.58	43.06	44.14	44.78	
46	04.51	08.97	13.35	17.60	21.68	25.56	29.18	32.53	35.56	38.25	40.56	42.50	44.02	45.12	45.78	
47	04.61	09.17	13.64	17.99	22.16	26.11	29.82	33.23	36.33	39.08	41.45	43.42	44.97	46.10	46.77	
48	04.70	09.36	13.93	18.37	22.63	26.67	30.45	33.94	37.10	39.91	42.33	44.35	45.93	47.08	47.77	
49	04.80	09.56	14.22	18.75	23.10	27.22	31.09	34.65	37.88	40.74	43.21	45.27	46.89	48.06	48.76	
50	04.90	09.75	14.52	19.14	23.57	27.78	31.72	35.36	38.65	41.57	44.10	46.20	47.85	49.04	49.76	
51	05.00	09.95	14.81	19.52	24.04	28.34	32.35	36.06	39.42	42.41	44.98	47.12	48.80	50.02	50.76	
52	05.10	10.14	15.10	19.90	24.51	28.89	32.99	36.77	40.20	43.24	45.86	48.04	49.76	51.00	51.75	
53	05.19	10.34	15.39	20.28	24.98	29.45	33.62	37.48	40.97	44.07	46.74	48.97	50.72	51.98	52.75	
54	05.29	10.53	15.68	20.67	25.46	30.00	34.26	38.18	41.74	44.90	47.62	49.89	51.67	52.96	53.74	
55	05.39	10.73	15.97	21.05	25.93	30.56	34.89	38.89	42.52	45.73	48.50	50.81	52.63	53.94	54.74	
56	05.49	10.92	16.26	21.43	26.40	31.11	35.53	39.60	43.29	46.56	49.39	51.74	53.59	54.92	55.73	
57	05.59	11.12	16.55	21.81	26.87	31.67	36.16	40.30	44.06	47.40	50.27	52.66	54.54	55.91	56.73	
58	05.68	11.31	16.84	22.20	27.34	32.22	36.80	41.01	44.83	48.23	51.15	53.59	55.50	56.89	57.72	
59	05.78	11.51	17.13	22.58	27.81	32.78	37.43	41.72	45.61	49.06	52.03	54.51	56.46	57.87	58.72	
60	05.88	11.70	17.42	22.96	28.28	33.34	38.06	42.43	46.38	49.89	52.91	55.43	57.41	58.85	59.71	
61	05.98	11.90	17.71	23.34	28.76	33.89	38.70	43.13	47.15	50.72	53.80	56.36	58.37	59.83	60.71	
62	06.08	12.09	18.00	23.73	29.23	34.45	39.33	43.84	47.93	51.55	54.68	57.28	59.33	60.81	61.70	
63	06.17	12.29	18.29	24.11	29.70	35.00	39.97	44.55	48.70	52.38	55.56	58.21	60.28	61.79	62.70	
64	06.27	12.48	18.58	24.49	30.17	35.56	40.66	45.25	49.47	53.22	56.44	59.13	61.24	62.77	63.69	
65	06.37	12.68	18.87	24.88	30.64	36.11	41.24	45.96	50.25	54.05	57.32	60.05	62.20	63.75	64.69	
66	06.47	12.87	19.16	25.26	31.11	36.67	41.87	46.67	51.02	54.88	58.21	60.98	63.16	64.73	65.68	
67	06.57	13.07	19.45	25.64	31.58	37.22	42.50	47.38	51.79	55.71	59.09	61.90	64.11	65.71	66.68	
68	06.66	13.26	19.74	26.02	32.06	37.78	43.14	48.08	52.56	56.54	59.97	62.83	65.07	66.69	67.67	
69	06.76	13.46	20.03	26.41	32.53	38.34	43.77	48.79	53.34	57.37	60.85	63.75	66.03	67.68	68.67	
70	06.86	13.65	20.32	26.79	33.00	38.89	44.41	49.50	54.11	58.21	61.73	64.67	66.98	68.66	69.66	
71	06.96	13.85	20.61	27.17	33.47	39.45	45.04	50.20	54.88	59.04	62.61	65.60	67.94	69.68	70.66	
72	07.06	14.04	20.90	27.55	33.94	40.00	45.08	50.91	55.66	59.87	63.50	66.52	68.90	70.62	71.65	
73	07.15	14.24	21.19	27.94	34.41	40.56	46.31	51.62	56.43	60.70	64.38	67.44	69.85	71.60	72.65	
74	07.25	14.43	21.48	28.32	34.88	41.11	46.95	52.33	67.20	61.53	65.26	68.37	70.81	72.58	73.64	
75	07.35	14.63	21.77	28.70	35.36	41.67	47.58	53.03	57.98	62.36	66.14	69.29	71.77	73.56	74.64	
76	07.45	14.62	22.[illegible]6	29.09	35.83	42.23	48.21	53.74	58.75	63.19	67.02	70.22	72.72	74.54	75.64	
77	07.55	15.02	22.35	29.47	36.30	42.78	48.85	54.45	59.52	64.03	67.91	71.14	73.68	75.52	76.63	
78	07.64	15.21	22.64	29.85	36.77	43.34	49.48	55.15	60.29	64.86	68.79	72.06	74.64	76.50	77.63	
79	07.74	15.41	22.93	30.23	37.24	43.89	50.12	55.86	61.07	65.69	69.67	72.99	75.60	77.48	78.62	
80	07.84	15.60	23.22	30.62	37.71	44.45	50.75	56.57	61.84	66.52	70.55	73.91	76.55	78.46	79.62	
81	07.94	15.80	23.51	31.00	38.18	45.00	51.39	57.28	62.61	67.35	71.43	74.84	77.51	79.44	80.61	
82	08.04	15.99	23.80	31.38	38.65	45.56	52.02	57.98	63.39	68.18	72.32	75.76	78.47	80.43	81.61	
83	08.13	16.10	24.09	31.76	39.13	46.11	52.66	58.69	64.16	69.01	73.20	76.68	79.42	81.41	82.60	
84	08.23	16.38	24.39	32.15	39.60	46.67	53.29	59.40	64.93	69.85	74.08	77.61	80.38	82.39	83.60	
85	08.33	16.58	24.68	32.53	40.07	47.23	53.92	60.10	65.71	70.68	74.96	78.53	81.34	83.37	84.59	
86	08.43	16.77	24.97	32.91	40.54	47.78	54.56	60.81	66.48	71.51	75.84	79.46	82.29	84.35	85.59	
87	08.53	16.97	25.26	33.29	41.01	48.34	55.19	61.52	67.25	72.34	76.73	80.38	83.25	85.33	86.58	
88	08.62	17.16	25.55	33.68	41.48	48.89	55.82	62.22	68.02	73.17	77.61	81.30	84.21	86.31	87.58	
89	08.72	17.36	25.84	34.06	41.95	49.45	56.46	62.93	68.80	74.00	78.49	82.23	85.16	87.29	88.57	
90	08.82	17.55	26.13	34.44	42.43	50.00	57.10	63.64	69.57	74.84	79.37	83.15	86.12	88.27	89.57	
91	08.92	17.75	26.42	34.83	42.90	50.56	57.73	64.35	70.34	75.67	80.25	84.07	87.08	89.25	90.56	
92	09.02	17.94	26.71	35.21	43.37	51.12	58.36	65.05	71.12	76.50	81.13	85.00	88.03	90.23	91.56	
93	09.11	18.14	27.00	35.59	43.84	51.67	59.00	65.76	71.89	77.33	82.02	85.92	88.99	91.21	92.55	
94	09.21	18.33	27.29	35.97	44.31	52.23	59.63	66.47	72.66	78.16	82.90	86.85	89.95	92.20	93.55	
95	09.31	18.53	27.58	36.36	44.78	52.78	60.27	67.17	73.44	78.99	83.78	87.77	90.91	93.18	94.54	
96	09.41	18.72	27.87	36.74	45.25	53.34	60.90	67.88	74.21	79.82	84.66	88.69	91.86	94.15	95.54	
97	09.51	18.92	28.16	37.12	45.73	53.89	61.54	68.59	74.98	80.66	85.54	89.62	92.82	95.14	96.53	
98	09.60	19.11	28.45	37.50	46.20	54.45	62.17	69.30	75.75	81.49	86.43	90.54	93.78	96.12	72.53	
99	09.70	19.31	28.74	37.89	46.67	55.00	62.81	70.00	76.53	82.32	87.31	91.47	94.73	97.10	98.52	

TABLE I

F=.625MC

B_N	N							
FOR C_{SIN}	1 15 *17 *31	2 14 *18 *30	3 13 *19 *29	4 12 *20 *28	5 11 *21 *27	6 10 *22 *26	7 9 *23 *25	8 *24
FOR C_{COS}	7 *9 *23 25	6 *10 *22 26	5 *11 *21 27	4 *12 *2C 28	3 *13 *19 29	2 *14 *18 30	1 *15 *17 31	0 *16
1	00.20	00.38	00.56	00.71	00.83	00.92	00.98	1
2	00.39	00.77	01.11	01.41	01.66	01.85	01.96	2
3	00.59	01.15	01.67	02.12	02.49	02.77	02.94	3
4	00.78	01.53	02.22	02.83	03.33	03.70	03.92	4
5	00.98	01.91	02.78	03.54	04.16	04.62	04.90	5
6	01.17	02.30	03.33	04.24	04.99	05.54	05.88	6
7	01.37	02.68	03.89	04.95	05.82	06.47	06.86	7
8	01.56	03.06	04.44	05.66	06.05	07.39	07.85	8
9	01.76	03.44	05.00	06.36	07.48	08.32	08.83	9
10	01.95	03.83	05.56	07.07	08.32	09.24	09.81	10
11	02.15	04.21	06.11	07.78	09.15	10.16	10.79	11
12	02.34	04.60	06.67	08.49	09.98	11.09	11.77	12
13	02.54	04.98	07.22	09.19	10.81	12.01	12.75	13
14	02.73	05.36	07.78	09.90	11.64	12.93	13.73	14
15	02.93	05.74	08.33	10.61	12.47	13.86	14.71	15
16	03.12	06.12	08.89	11.31	13.30	14.78	15.69	16
17	03.32	06.51	09.45	12.02	14.14	15.71	16.67	17
18	03.51	06.89	10.00	12.73	14.97	16.63	17.65	18
19	03.71	07.27	10.56	13.43	15.80	17.55	18.64	19
20	03.90	07.65	11.11	14.14	16.63	18.48	19.62	20
21	04.10	08.04	11.67	14.85	17.46	19.40	20.60	21
22	04.29	08.42	12.22	15.56	18.29	20.33	21.58	22
23	04.49	08.80	12.78	16.26	19.12	21.25	22.56	23
24	04.68	09.18	13.33	16.97	19.96	22.17	23.54	24
25	04.88	09.57	13.85	17.68	20.79	23.10	24.52	25
26	05.07	09.95	14.45	18.38	21.62	24.02	25.50	26
27	05.27	10.33	15.00	19.09	22.45	24.95	26.48	27
28	05.46	10.72	15.56	19.80	23.28	25.87	27.46	28
29	05.66	11.10	16.11	20.51	24.11	26.79	28.44	29
30	05.85	11.48	16.67	21.21	24.95	27.72	29.42	30
31	06.05	11.86	17.22	21.92	25.78	28.64	30.40	31
32	06.24	12.25	17.78	22.63	26.61	29.56	31.39	32
33	06.44	12.63	18.33	23.33	27.44	30.49	32.37	33
34	06.63	13.01	18.89	24.04	28.27	31.41	33.35	34
35	06.83	13.39	19.45	24.75	29.10	32.34	34.33	35
36	07.02	13.78	20.00	25.46	29.93	33.26	35.31	36
37	07.22	14.16	20.56	26.16	30.77	34.18	36.29	37
38	07.41	14.54	21.11	26.87	31.60	35.11	37.27	38
39	07.61	14.92	21.67	27.58	32.43	36.03	38.25	39
40	07.80	15.31	22.22	28.28	33.26	36.96	39.23	40
41	08.00	15.69	22.78	28.99	34.09	37.88	40.21	41
42	08.19	16.07	23.34	29.70	34.92	38.80	41.19	42
43	08.39	16.46	23.89	30.41	35.75	39.73	42.17	43
44	08.58	16.84	24.45	31.11	36.59	40.65	43.16	44
45	08.78	17.22	25.00	31.82	37.42	41.58	44.14	45
46	08.97	17.60	25.56	32.53	38.25	42.50	45.12	46
47	09.17	17.99	26.11	33.23	39.08	43.42	46.10	47
48	09.36	18.37	26.67	33.94	39.91	44.35	47.08	48
49	09.56	18.75	27.22	34.65	40.74	45.27	48.06	49
50	09.75	19.14	27.78	35.36	41.57	46.20	49.04	50
51	09.95	19.52	28.34	36.06	42.41	47.12	50.02	51
52	10.14	19.90	28.89	36.77	43.24	48.04	51.00	52
53	10.34	20.28	29.45	37.48	44.07	48.97	51.98	53
54	10.53	20.67	30.00	38.18	44.90	49.89	52.96	54
55	10.73	21.05	30.56	38.89	45.73	50.81	53.94	55
56	10.92	21.43	31.11	39.60	46.56	51.74	54.92	56
57	11.12	21.81	31.67	40.30	47.40	52.66	55.91	57
58	11.31	22.20	32.22	41.01	48.23	53.59	56.89	58
59	11.51	22.58	32.78	41.72	49.06	54.51	57.87	59
60	11.70	22.96	33.34	42.43	49.89	55.43	58.85	60
61	11.90	23.34	33.89	43.13	50.72	56.36	59.83	61
62	12.09	23.73	34.45	43.84	51.55	57.28	60.81	62
63	12.29	24.11	35.00	44.55	52.38	58.21	61.79	63
64	12.48	24.49	35.56	45.25	53.22	59.13	62.77	64
65	12.68	24.88	36.11	45.96	54.05	50.05	63.75	65
66	12.87	25.26	36.67	46.67	54.88	60.98	64.73	66
67	13.07	25.64	37.22	47.38	55.71	61.90	65.71	67
68	13.26	26.02	37.78	48.08	56.54	62.83	66.69	68
69	13.46	26.41	38.34	48.79	57.37	63.75	67.68	69
70	13.65	26.79	38.89	49.50	58.21	54.67	68.66	70
71	13.85	27.17	39.45	50.20	59.04	65.60	69.68	71
72	14.04	27.55	40.00	50.91	59.87	66.52	70.62	72
73	14.24	27.94	40.56	51.62	60.70	67.44	71.60	73
74	14.43	28.32	41.11	52.33	61.53	68.37	72.58	74
75	14.63	28.70	41.67	53.03	62.36	69.29	73.56	75
76	14.82	29.09	42.23	53.74	53.19	70.22	74.54	76
77	15.02	29.47	42.78	54.45	54.03	71.14	75.52	77
78	15.21	29.85	43.34	55.15	54.86	72.06	76.50	78
79	15.41	30.23	43.89	55.86	65.69	72.99	77.48	79
80	15.60	30.62	44.45	56.57	56.52	13.91	78.46	80
81	15.80	31.00	45.00	57.28	67.35	74.84	79.44	81
82	15.99	31.38	45.56	57.98	68.18	75.76	80.43	82
83	16.10	31.76	46.11	58.69	59.01	76.68	81.41	83
84	16.38	32.15	46.67	59.40	59.85	77.61	82.39	84
85	16.58	32.53	47.23	60.10	70.68	78.53	83.37	85
86	16.77	32.91	47.78	60.81	71.51	79.46	84.35	86
87	16.97	33.29	48.34	61.52	72.34	80.38	85.33	87
88	17.16	33.68	48.89	62.22	73.17	81.30	86.31	88
89	17.36	34.06	49.45	62.93	74.00	82.23	87.29	89
90	17.55	34.44	50.00	63.64	74.84	83.15	88.27	90
91	17.75	34.83	50.56	64.35	75.67	84.07	89.25	91
92	17.94	35.21	51.12	65.05	76.50	85.00	90.23	92
93	18.14	35.59	51.67	65.76	77.33	85.92	91.21	93
94	18.33	35.97	52.23	66.47	78.16	86.85	92.20	94
95	18.53	36.36	52.78	67.17	78.99	87.77	93.18	95
96	18.72	36.74	53.34	67.88	79.82	88.69	94.15	96
97	18.92	37.12	53.89	68.59	30.66	89.62	95.14	97
98	19.11	37.50	54.45	69.30	31.49	90.54	96.12	98
99	19.31	37.89	55.00	70.00	32.32	91.47	97.10	99

TABLE II

F=1.25MC

B_N	N			
FOR C_{SIN}	1 7 *9 *15 17 23 *25 *31	2 6 *10 *14 18 22 *26 *30	3 5 *11 *13 19 21 *27 *29	4 *12 20 *28
FOR C_{COS}	3 *5 *11 13 19 *21 *27 29	2 *6 *10 14 18 *22 *26 30	1 *7 *9 15 17 *23 *25 31	0 *8 16 *24
1	00.38	00.71	00.92	1
2	00.77	01.41	01.85	2
3	01.15	02.12	02.77	3
4	01.53	02.83	03.70	4
5	01.91	03.54	04.62	5
6	02.30	04.24	05.54	6
7	02.68	04.95	06.47	7
8	03.06	05.66	07.39	8
9	03.44	06.36	08.32	9
10	03.83	07.07	09.24	10
11	04.21	07.78	10.16	11
12	04.60	08.49	11.09	12
13	04.98	09.19	12.01	13
14	05.36	09.90	12.93	14
15	05.74	10.61	13.86	15
16	06.12	11.31	14.78	16
17	06.51	12.02	15.71	17
18	06.89	12.73	16.63	18
19	07.27	13.43	17.55	19
20	07.65	14.14	18.48	20
21	08.04	14.85	19.40	21
22	08.42	15.56	20.33	22
23	08.80	16.26	21.25	23
24	09.18	16.97	22.17	24
25	09.57	17.68	23.10	25
26	09.95	18.38	24.02	26
27	10.33	19.09	24.95	27
28	10.72	19.80	25.87	28
29	11.10	20.51	26.79	29
30	11.48	21.21	27.72	30
31	11.86	21.92	28.64	31
32	12.25	22.63	29.56	32
33	12.63	23.33	30.49	33
34	13.01	24.04	31.41	34
35	13.39	24.75	32.34	35
36	13.78	25.46	33.26	36
37	14.16	26.16	34.18	37
38	14.54	26.87	35.11	38
39	14.92	27.58	36.03	39
40	15.31	28.28	36.96	40
41	15.69	28.99	37.88	41
42	16.07	29.70	38.80	42
43	16.46	30.41	39.73	43
44	16.84	31.11	40.65	44
45	17.22	31.82	41.58	45
46	17.60	32.53	42.50	46
47	17.99	33.23	43.42	47
48	18.37	33.94	44.35	48
49	18.75	34.65	45.27	49
50	19.14	35.36	46.20	50
51	19.52	36.06	47.12	51
52	19.90	36.77	48.04	52
53	20.28	37.48	48.97	53
54	20.67	38.18	49.89	54
55	21.05	38.89	50.81	55
56	21.43	39.60	51.74	56
57	21.81	40.30	52.66	57
58	22.20	41.01	53.59	58
59	22.58	41.72	54.51	59
60	22.96	42.43	55.43	60
61	23.34	43.13	56.36	61
62	23.73	43.84	57.28	62
63	24.11	44.55	58.21	63
64	24.49	45.25	59.13	64
65	24.88	45.96	60.05	65
66	25.26	46.67	60.98	66
67	25.64	47.38	61.90	67
68	26.02	48.08	62.83	68
69	26.41	48.79	63.75	69
70	26.79	49.50	64.67	70
71	27.17	50.20	65.60	71
72	27.55	50.91	66.52	72
73	27.94	51.62	67.44	73
74	28.32	52.33	68.37	74
75	28.70	53.03	69.29	75
76	29.09	53.74	70.22	76
77	29.47	54.45	71.14	77
78	29.85	55.15	72.06	78
79	30.23	55.86	72.99	79
80	30.62	56.57	73.91	80
81	31.00	57.28	74.84	81
82	31.38	57.98	75.76	82
83	31.76	58.69	76.68	83
84	32.15	59.40	77.61	84
85	32.53	60.10	78.53	85
86	32.91	60.81	79.46	86
87	33.29	61.52	80.38	87
88	33.68	62.22	81.30	88
89	34.06	62.93	82.23	89
90	34.44	63.64	83.15	90
91	34.83	64.35	84.07	91
92	35.21	65.05	85.00	92
93	35.59	65.76	85.92	93
94	35.97	66.47	86.85	94
95	36.36	67.17	87.77	95
96	36.74	67.88	88.69	96
97	37.12	68.59	89.62	97
98	37.50	69.30	90.54	98
99	37.89	70.00	91.47	99

TABLE III

F=1.875MC

B_N	N: 5, 11, *21, *27 (for C_{SIN}); *3, 13, 19, *29 (for C_{COS})	N: *6, *10, 22, 26 (for C_{SIN}); 2, *14, *18, 30 (for C_{COS})	N: 1, 15, *17, *31 (for C_{SIN}); *7, 9, 23, *25 (for C_{COS})	N: 4, 12, *20, *28 (for C_{SIN}); *4, 12, 20, *28 (for C_{COS})	N: *7, *9, 23, 25 (for C_{SIN}); 1, *15, *17, 31 (for C_{COS})	N: 2, 14, *18, *30 (for C_{SIN}); *6, 10, 22, *26 (for C_{COS})	N: 3, 13, *19, *29 (for C_{SIN}); *5, 11, 21, *27 (for C_{COS})	N: *8, 24 (for C_{SIN}); 0, *16 (for C_{COS})
1	00.20	00.38	00.56	00.71	00.83	00.92	00.98	1
2	00.39	00.77	01.11	01.41	01.66	01.85	01.96	2
3	00.59	01.15	01.67	02.12	02.49	02.77	02.94	3
4	00.78	01.53	02.22	02.83	03.33	03.70	03.92	4
5	00.98	01.91	02.78	03.54	04.16	04.62	04.90	5
6	01.17	02.30	03.33	04.24	04.99	05.54	05.88	6
7	01.37	02.68	03.89	04.95	05.82	06.47	06.86	7
8	01.56	03.06	04.44	05.66	06.05	07.39	07.85	8
9	01.76	03.44	05.00	06.36	07.48	08.32	08.83	9
10	01.95	03.83	05.56	07.07	08.32	09.24	09.81	10
11	02.15	04.21	06.11	07.78	09.15	10.16	10.79	11
12	02.34	04.60	06.67	08.49	09.98	11.09	11.77	12
13	02.54	04.98	07.22	09.19	10.81	12.01	12.75	13
14	02.73	05.36	07.78	09.90	11.64	12.93	13.73	14
15	02.93	05.74	08.33	10.61	12.47	13.86	14.71	15
16	03.12	06.12	08.89	11.31	13.30	14.78	15.69	16
17	03.32	06.51	09.45	12.02	14.14	15.71	16.67	17
18	03.51	06.89	10.00	12.73	14.97	16.63	17.65	18
19	03.71	07.27	10.56	13.43	15.80	17.55	18.64	19
20	03.90	07.65	11.11	14.14	16.63	18.48	19.62	20
21	04.10	08.04	11.67	14.85	17.46	19.40	20.60	21
22	04.29	08.42	12.22	15.56	18.29	20.33	21.58	22
23	04.49	08.80	12.78	16.26	19.12	21.25	22.56	23
24	04.68	09.18	13.33	16.97	19.96	22.17	23.54	24
25	04.88	09.57	13.89	17.68	20.79	23.10	24.52	25
26	05.07	09.95	14.45	18.38	21.62	24.02	25.50	26
27	05.27	10.33	15.00	19.09	22.45	24.95	26.48	27
28	05.46	10.72	15.56	19.80	23.28	25.87	27.46	28
29	05.66	11.10	16.11	20.51	24.11	26.79	28.44	29
30	05.85	11.48	16.67	21.21	24.95	27.72	29.42	30
31	06.05	11.86	17.22	21.92	25.78	28.64	30.40	31
32	06.24	12.25	17.78	22.63	26.61	29.56	31.39	32
33	06.44	12.63	18.33	23.33	27.44	30.49	32.37	33
34	06.63	13.01	18.89	24.04	28.27	31.41	33.35	34
35	06.83	13.39	19.45	24.75	29.10	32.34	34.33	35
36	07.02	13.78	20.00	25.46	29.93	33.26	35.31	36
37	07.22	14.16	20.56	26.16	30.77	34.18	36.29	37
38	07.41	14.54	21.11	26.87	31.60	35.11	37.27	38
39	07.61	14.92	21.67	27.58	32.43	36.03	38.25	39
40	07.80	15.31	22.22	28.28	33.26	36.96	39.23	40
41	08.00	15.69	22.78	28.99	34.09	37.88	40.21	41
42	08.19	16.07	23.34	29.70	34.92	38.80	41.19	42
43	08.39	16.46	23.89	30.41	35.75	39.73	42.17	43
44	08.58	16.84	24.45	31.11	36.59	40.65	43.16	44
45	08.78	17.22	25.00	31.82	37.42	41.58	44.14	45
46	08.97	17.60	25.56	32.53	38.25	42.50	45.12	46
47	09.17	17.99	26.11	33.23	39.08	43.42	46.10	47
48	09.36	18.37	26.67	33.94	39.91	44.35	47.08	48
49	09.56	18.75	27.22	34.65	40.74	45.27	48.06	49
50	09.75	19.14	27.78	35.36	41.57	46.20	49.04	50
51	09.95	19.52	28.34	36.06	42.41	47.12	50.02	51
52	10.14	19.90	28.89	36.77	43.24	48.04	51.00	52
53	10.34	20.28	29.45	37.48	44.07	48.97	51.98	53
54	10.53	20.67	30.00	38.18	44.90	49.89	52.96	54
55	10.73	21.05	30.56	38.89	45.73	50.81	53.94	55
56	10.92	21.43	31.11	39.60	46.56	51.74	54.92	56
57	11.12	21.81	31.67	40.30	47.40	52.66	55.91	57
58	11.31	22.20	32.22	41.01	48.23	53.59	56.89	58
59	11.51	22.58	32.78	41.72	49.06	54.51	57.87	59
60	11.70	22.96	33.34	42.43	49.89	55.43	58.85	60
61	11.90	23.34	33.89	43.13	50.72	56.36	59.83	61
62	12.09	23.73	34.45	43.84	51.55	57.28	60.81	62
63	12.29	24.11	35.00	44.55	52.38	58.21	61.79	63
64	12.48	24.49	35.56	45.25	53.22	59.13	62.77	64
65	12.68	24.88	36.11	45.96	54.05	60.05	63.75	65
66	12.87	25.26	36.67	46.67	54.88	60.98	64.73	66
67	13.07	25.64	37.22	47.38	55.71	61.90	65.71	67
68	13.26	26.02	37.78	48.08	56.54	62.83	66.69	68
69	13.46	26.41	38.34	48.79	57.37	63.75	67.68	69
70	13.65	26.79	38.89	49.50	58.21	64.67	68.66	70
71	13.85	27.17	39.45	50.20	59.04	65.60	69.64	71
72	14.04	27.55	40.00	50.91	59.87	66.52	70.62	72
73	14.24	27.94	40.56	51.62	60.70	67.44	71.60	73
74	14.43	28.32	41.11	52.33	61.53	68.37	72.58	74
75	14.63	28.70	41.67	53.03	62.36	69.29	73.56	75
76	14.82	29.09	42.23	53.74	63.19	70.22	74.54	76
77	15.02	29.47	42.78	54.45	64.03	71.14	75.52	77
78	15.21	29.85	43.34	55.15	64.86	72.06	76.50	78
79	15.41	30.23	43.89	55.86	65.69	72.99	77.48	79
80	15.60	30.62	44.45	56.57	66.52	73.91	78.46	80
81	15.80	31.00	45.00	57.28	67.35	74.84	79.44	81
82	15.99	31.38	45.56	57.98	68.18	75.76	80.43	82
83	16.10	31.76	46.11	58.69	69.01	76.68	81.41	83
84	16.38	32.15	46.67	59.40	69.85	77.61	82.39	84
85	16.58	32.53	47.23	60.10	70.68	78.53	83.37	85
86	16.77	32.91	47.78	60.81	71.51	79.46	84.35	86
87	16.97	33.29	48.34	61.52	72.34	80.38	85.33	87
88	17.16	33.68	48.89	62.22	73.17	81.30	86.31	88
89	17.36	34.06	49.45	62.93	74.00	82.23	87.29	89
90	17.55	34.44	50.00	63.64	74.84	83.15	88.27	90
91	17.75	34.83	50.56	64.35	75.67	84.07	89.25	91
92	17.94	35.21	51.12	65.05	76.50	85.00	90.23	92
93	18.14	35.59	51.67	65.76	77.33	85.92	91.21	93
94	18.33	35.97	52.23	66.47	78.16	86.85	92.20	94
95	18.53	36.36	52.78	67.17	78.99	87.77	93.18	95
96	18.72	36.74	53.34	67.88	79.82	88.69	94.15	96
97	18.92	37.12	53.89	68.59	80.66	89.62	95.14	97
98	19.11	37.50	54.45	69.30	81.49	90.54	96.12	98
99	19.31	37.89	55.00	70.00	82.32	91.47	97.10	99

TABLE IV

F=2.5MC

B_N	N: 1, 3, *5, *7, 9, 11, *13, *15, 17, 19, *21, *23, 25, 27, *29, *31 (for C_{SIN}); 1, *3, *5, 7, 9, *11, *13, 15, 17, *19, *21, 23, 25, *27, *29, 31 (for C_{COS})	N: 2, *6, 10, *14, 18, *22, 26, *30 (for C_{SIN}); 0, *4, 8, *12, 16, *20, 24, *28 (for C_{COS})
1	00.71	1
2	01.41	2
3	02.12	3
4	02.83	4
5	03.54	5
6	04.24	6
7	04.95	7
8	05.66	8
9	06.36	9
10	07.07	10
11	07.78	11
12	08.49	12
13	09.19	13
14	09.90	14
15	10.61	15
16	11.31	16
17	12.02	17
18	12.73	18
19	13.43	19
20	14.14	20
21	14.85	21
22	15.56	22
23	16.26	23
24	16.97	24
25	17.68	25
26	18.38	26
27	19.09	27
28	19.80	28
29	20.51	29
30	21.21	30
31	21.92	31
32	22.63	32
33	23.33	33
34	24.04	34
35	24.75	35
36	25.46	36
37	26.16	37
38	26.87	38
39	27.58	39
40	28.28	40
41	28.99	41
42	29.70	42
43	30.41	43
44	31.11	44
45	31.82	45
46	32.53	46
47	33.23	47
48	33.94	48
49	34.65	49
50	35.36	50
51	36.06	51
52	36.77	52
53	37.48	53
54	38.18	54
55	38.89	55
56	39.60	56
57	40.30	57
58	41.01	58
59	41.72	59
60	42.43	60
61	43.13	61
62	43.84	62
63	44.55	63
64	45.25	64
65	45.96	65
66	46.67	66
67	47.38	67
68	48.08	68
69	48.79	69
70	49.50	70
71	50.20	71
72	50.91	72
73	51.62	73
74	52.33	74
75	53.03	75
76	53.74	76
77	54.45	77
78	55.15	78
79	55.86	79
80	56.57	80
81	57.28	81
82	57.98	82
83	58.69	83
84	59.40	84
85	60.10	85
86	60.81	86
87	61.52	87
88	62.22	88
89	62.93	89
90	63.64	90
91	64.35	91
92	65.05	92
93	65.76	93
94	66.47	94
95	67.17	95
96	67.88	96
97	68.59	97
98	69.30	98
99	70.00	99

TABLE V

F=3.125MC

B_N	N: 3 13 *19 *29 (for C_{SIN}) / 5 *11 *21 27 (for C_{COS})	*6 *10 22 26 / *2 14 18 *30	7 9 *23 *25 / 1 *15 *17 31	*4 *12 20 28 / *4 12 20 *28	1 15 *17 *31 / 7 *9 *23 25	2 14 *18 *30 / 6 *10 *22 26	*5 *11 21 27 / *3 13 19 *29	8 *24 / 0 *16
1	00.20	00.38	00.56	00.71	00.83	00.92	00.98	1
2	00.39	00.77	01.11	01.41	01.66	01.85	01.96	2
3	00.59	01.15	01.67	02.12	02.49	02.77	02.94	3
4	00.78	01.53	02.22	02.83	03.33	03.70	03.92	4
5	00.98	01.91	02.78	03.54	04.16	04.62	04.90	5
6	01.17	02.30	03.33	04.24	04.99	05.54	05.88	6
7	01.37	02.68	03.89	04.95	05.82	06.47	06.86	7
8	01.56	03.06	04.44	05.66	06.05	07.39	07.85	8
9	01.76	03.44	05.00	06.36	07.48	08.32	08.83	9
10	01.95	03.83	05.56	07.07	08.32	09.24	09.81	10
11	02.15	04.21	06.11	07.78	09.15	10.16	10.79	11
12	02.34	04.60	06.67	08.49	09.98	11.09	11.77	12
13	02.54	04.98	07.22	09.19	10.81	12.01	12.75	13
14	02.73	05.36	07.78	09.90	11.64	12.93	13.73	14
15	02.93	05.74	08.33	10.61	12.47	13.86	14.71	15
16	03.12	06.12	08.89	11.31	13.30	14.78	15.69	16
17	03.32	06.51	09.45	12.02	14.14	15.71	16.67	17
18	03.51	06.89	10.00	12.73	14.97	16.63	17.65	18
19	03.71	07.27	10.56	13.43	15.80	17.55	18.64	19
20	03.90	07.65	11.11	14.14	16.63	18.48	19.62	20
21	04.10	08.04	11.67	14.85	17.46	19.40	20.60	21
22	04.29	08.42	12.22	15.56	18.29	20.33	21.58	22
23	04.49	08.80	12.78	16.26	19.12	21.25	22.56	23
24	04.68	09.18	13.33	16.97	19.96	22.17	23.54	24
25	04.88	09.57	13.89	17.68	20.79	23.10	24.52	25
26	05.07	09.95	14.45	18.38	21.62	24.02	25.50	26
27	05.27	10.33	15.00	19.09	22.45	24.95	26.48	27
28	05.46	10.72	15.56	19.80	23.28	25.87	27.46	28
29	05.66	11.10	16.11	20.51	24.11	26.79	28.44	29
30	05.85	11.48	16.67	21.21	24.95	27.72	29.42	30
31	06.05	11.86	17.22	21.92	25.78	28.64	30.40	31
32	06.24	12.25	17.78	22.63	26.61	29.56	31.39	32
33	06.44	12.63	18.33	23.33	27.44	30.49	32.37	33
34	06.63	13.01	18.89	24.04	28.27	31.41	33.35	34
35	06.83	13.39	19.45	24.75	29.10	32.34	34.33	35
36	07.02	13.78	20.00	25.46	29.93	33.26	35.31	36
37	07.22	14.16	20.56	26.16	30.77	34.18	36.29	37
38	07.41	14.54	21.11	26.87	31.60	35.11	37.27	38
39	07.61	14.92	21.67	27.58	32.43	36.03	38.25	39
40	07.80	15.31	22.22	28.28	33.26	36.96	39.23	40
41	08.00	15.69	22.78	28.99	34.09	37.88	40.21	41
42	08.19	16.07	23.34	29.70	34.92	38.80	41.19	42
43	08.39	16.46	23.89	30.41	35.75	39.73	42.17	43
44	08.58	16.84	24.45	31.11	36.59	40.65	43.16	44
45	08.78	17.22	25.00	31.82	37.42	41.58	44.14	45
46	08.97	17.60	25.56	32.53	38.25	42.50	45.12	46
47	09.17	17.99	26.11	33.23	39.08	43.42	46.10	47
48	09.36	18.37	26.67	33.94	39.91	44.35	47.08	48
49	09.56	18.75	27.22	34.65	40.74	45.27	48.06	49
50	09.75	19.14	27.78	35.36	41.57	46.20	49.04	50
51	09.95	19.52	28.34	36.06	42.41	47.12	50.02	51
52	10.14	19.90	28.89	36.77	43.24	48.04	51.00	52
53	10.34	20.28	29.45	37.48	44.07	48.97	51.98	53
54	10.53	20.67	30.00	38.18	44.90	49.89	52.96	54
55	10.73	21.05	30.56	38.89	45.73	50.81	53.94	55
56	10.92	21.43	31.11	39.60	46.56	51.74	54.92	56
57	11.12	21.81	31.67	40.30	47.40	52.66	55.91	57
58	11.31	22.20	32.22	41.01	48.23	53.59	56.89	58
59	11.51	22.58	32.78	41.72	49.06	54.51	57.87	59
60	11.70	22.96	33.34	42.43	49.89	55.43	58.85	60
61	11.90	23.34	33.89	43.13	50.72	56.36	59.83	61
62	12.09	23.73	34.45	43.84	51.55	57.28	60.81	62
63	12.29	24.11	35.00	44.55	52.38	58.21	61.79	63
64	12.48	24.49	35.56	45.25	53.22	59.13	62.77	64
65	12.68	24.88	36.11	45.96	54.05	60.05	63.75	65
66	12.87	25.26	36.67	46.67	54.88	60.98	64.73	66
67	13.07	25.64	37.22	47.38	55.71	61.90	65.71	67
68	13.26	26.02	37.78	48.08	56.54	62.83	66.69	68
69	13.46	26.41	38.34	48.79	57.37	63.75	67.68	69
70	13.65	26.79	38.89	49.50	58.21	64.67	68.66	70
71	13.85	27.17	39.45	50.20	59.04	65.60	69.64	71
72	14.04	27.55	40.00	50.91	59.87	66.52	70.62	72
73	14.24	27.94	40.56	51.62	60.70	67.44	71.60	73
74	14.43	28.32	41.11	52.33	61.53	68.37	72.58	74
75	14.63	28.70	41.67	53.03	62.36	69.29	73.56	75
76	14.62	29.09	42.23	53.74	53.19	70.22	74.54	76
77	15.02	29.47	42.78	54.45	54.03	71.14	75.52	77
78	15.21	29.85	43.34	55.15	54.86	72.06	76.50	78
79	15.41	30.23	43.89	55.86	65.69	72.99	77.48	79
80	15.60	30.62	44.45	56.57	66.52	73.91	78.46	80
81	15.80	31.00	45.00	57.28	67.35	74.84	79.44	81
82	15.99	31.38	45.56	57.98	68.18	75.76	80.43	82
83	16.10	31.76	46.11	58.69	69.01	76.68	81.41	83
84	16.38	32.15	46.67	59.40	69.85	77.61	82.39	84
85	16.58	32.53	47.23	60.10	70.68	78.53	83.37	85
86	16.77	32.91	47.78	60.81	71.51	79.46	84.35	86
87	16.97	33.29	48.34	61.52	72.34	80.38	85.33	87
88	17.16	33.68	48.89	62.22	73.17	81.30	86.31	88
89	17.36	34.06	49.45	62.93	74.00	82.23	87.29	39
90	17.55	34.44	50.00	63.64	74.84	83.15	88.27	90
91	17.75	34.83	50.56	64.35	75.67	84.07	89.25	91
92	17.94	35.21	51.12	65.05	76.50	85.00	90.23	92
93	18.14	35.59	51.67	65.76	77.33	85.92	91.21	93
94	18.33	35.97	52.23	66.47	78.16	86.85	92.20	94
95	18.53	36.36	52.78	67.17	78.99	87.77	93.18	95
96	18.72	36.74	53.34	67.88	79.82	88.69	94.15	96
97	18.92	37.12	53.89	68.59	30.66	89.62	95.14	97
98	19.11	37.50	54.45	69.30	81.49	90.54	96.12	98
99	19.31	37.89	55.00	70.00	82.32	91.47	97.10	99

TABLE VI

F=3.75MC

B_N	N: *3 *5 11 13 *19 *21 27 29 (for C_{SIN}) / 1 *7 *9 15 17 *23 *25 31 (for C_{COS})	2 6 *10 *14 18 22 *26 *30 / *2 6 10 *14 *18 22 26 *30	1 7 *9 *15 17 23 *25 *31 / *3 5 11 *13 *19 21 27 *29	*4 12 *20 28 / 0 *8 16 *24
1	00.38	00.71	00.92	1
2	00.77	01.41	01.85	2
3	01.15	02.12	02.77	3
4	01.53	02.83	03.70	4
5	01.91	03.54	04.62	5
6	02.30	04.24	05.54	6
7	02.68	04.95	06.47	7
8	03.06	05.66	07.39	8
9	03.44	06.36	08.32	9
10	03.83	07.07	09.24	10
11	04.21	07.78	10.16	11
12	04.60	08.49	11.09	12
13	04.98	09.19	12.01	13
14	05.36	09.90	12.93	14
15	05.74	10.61	13.86	15
16	06.12	11.31	14.78	16
17	06.51	12.02	15.71	17
18	06.89	12.73	16.63	18
19	07.27	13.43	17.55	19
20	07.65	14.14	18.48	20
21	08.04	14.85	19.40	21
22	08.42	15.56	20.33	22
23	08.80	16.26	21.25	23
24	09.18	16.97	22.17	24
25	09.57	17.68	23.10	25
26	09.95	18.38	24.02	26
27	10.33	19.09	24.95	27
28	10.72	19.80	25.87	28
29	11.10	20.51	26.79	29
30	11.48	21.21	27.72	30
31	11.86	21.92	28.64	31
32	12.25	22.63	29.56	32
33	12.63	23.33	30.49	33
34	13.01	24.04	31.41	34
35	13.39	24.75	32.34	35
36	13.78	25.46	33.26	36
37	14.16	26.16	34.18	37
38	14.54	26.87	35.11	38
39	14.92	27.58	36.03	39
40	15.31	28.28	36.96	40
41	15.69	28.99	37.88	41
42	16.07	29.70	38.80	42
43	16.46	30.41	39.73	43
44	16.84	31.11	40.65	44
45	17.22	31.82	41.58	45
46	17.60	32.53	42.50	46
47	17.99	33.23	43.42	47
48	18.37	33.94	44.35	48
49	18.75	34.65	45.27	49
50	19.14	35.36	46.20	50
51	19.52	36.06	47.12	51
52	19.90	36.77	48.04	52
53	20.28	37.48	48.97	53
54	20.67	38.18	49.89	54
55	21.05	38.89	50.81	55
56	21.43	39.60	51.74	56
57	21.81	40.30	52.66	57
58	22.20	41.01	53.59	58
59	22.58	41.72	54.51	59
60	22.96	42.43	55.43	60
61	23.34	43.13	56.36	61
62	23.73	43.84	57.28	62
63	24.11	44.55	58.21	63
64	24.49	45.25	59.13	64
65	24.88	45.96	60.05	65
66	25.26	46.67	60.98	66
67	25.64	47.38	61.90	67
68	26.02	48.08	62.83	68
69	26.41	48.79	63.75	69
70	26.79	49.50	64.67	70
71	27.17	50.20	65.60	71
72	27.55	50.91	66.52	72
73	27.94	51.62	67.44	73
74	28.32	52.33	68.37	74
75	28.70	53.03	69.29	75
76	29.09	53.74	70.22	76
77	29.47	54.45	71.14	77
78	29.85	55.15	72.06	78
79	30.23	55.86	72.99	79
80	30.62	56.57	73.91	80
81	31.00	57.28	74.84	81
82	31.38	57.98	75.76	82
83	31.16	58.69	76.68	83
84	32.15	59.40	77.61	84
85	32.53	60.10	78.53	85
86	32.91	60.81	79.46	86
87	33.29	61.52	80.38	87
88	33.68	62.22	81.30	88
89	34.06	62.93	82.23	89
90	34.44	63.64	83.15	90
91	34.83	64.35	84.07	91
92	35.21	65.05	85.00	92
93	35.59	65.76	85.92	93
94	35.97	66.47	86.85	94
95	36.36	67.17	87.77	95
96	36.74	67.88	88.69	96
97	37.12	68.59	89.62	97
98	37.50	69.30	90.54	98
99	37.89	70.00	91.47	99

TABLE VII

TABLE VIII

$F=5$ *Mc*

For C_{sin}

Add the values of B_n for the following subscripts (retaining proper signs)

1 5 9 13 17 21 25 29

and add the values for the following subscripts (reversing their signs)

3 7 11 15 19 23 27 31

For C_{cos}

Add the values of B_n for the following subscripts (retaining proper signs)

0 4 8 12 16 20 24 28

and add the values for the following subscripts (reversing their signs)

2 6 10 14 18 22 26 30

For the calculation of the auxiliary value C_{cos} proceed in exactly the same way, using the heading "for C_{cos}." The tables are prepared for a maximum of 32 points. It is, however, easy to use these tables for any number of points if the following is kept in mind:

> For all frequencies except the lowest (0.3125 Mc) the column which applies to B_n applies as well to B_{n+32}. The same is true for the lowest frequency (0.3125 Mc) but, in addition, the sign of B_{n+32} must be reversed.

Computation of Amplitude

Use nomograph II unless the first digit of the larger one of the auxiliary values C_{sin} and C_{cos} lies between 2 and 8, in which case nomograph I should be used for higher accuracy. Connect, with the aid of a straight edge, the C_{sin} value on scale 1 with the C_{cos} value of scale 2. The value at the intersection of scale 3 with the straight edge is the desired value in cases where the frequency is equal to or smaller than 1.875 Mc. For higher frequencies, proceed perpendicularly from the value found on scale 3 to the scale marked with the proper frequency.

Computation of Phase

Use nomograph III. Connect the values C_{sin} on scale 1 with the value of C_{cos} on scale 2 with a straight edge and find the phase angle α on scale 3 at the intersection with the straight edge. If C_{sin} and C_{cos} have the same sign, take the angle as positive; if they have opposite signs, take it as negative.

Example:

Subscript n of sample points	0	1	2	3	4	5	6	7
A_n	1	9.2	21.9	17.5	18·5	18.0	18.5	18.0
B_n	1	8.2	12.7	−4.4	1	−0.5	0.5	−5
$B_n \times$ Scale Factor (=7)	7	57	89	−31	7	−4	4	−4

Evaluation for $f_0=0.625$ Mc/s

$C_{sin}=11.12+34.0-17.22+4.95-3.33+3.7-3.92 \quad = 29.3$

$C_{cos}=-0.78+1.53-2.22+4.95-25.78+82.23+55.91+7=122.84$

Nomograph II gives $|H(0.625 \text{ Mc})|=126.0$

Nomograph III gives $\alpha \quad = 13.5°$

F=6.25 MC

B_N	N				
	*3 *5 11 13 *19 *21 27 29	*2 *6 10 14 *18 *22 26 30	1 7 *9 *15 17 23 *25 *31	4 *12 20 *28	FOR C_{SIN}
	*1 7 9 *15 *17 23 25 *31	*2 6 10 *14 *18 22 26 *30	3 *5 *11 13 19 *21 *27 29	0 *8 16 *24	FOR C_{COS}
1	00.38	00.71	00.92	1	
2	00.77	01.41	01.85	2	
3	01.15	02.12	02.77	3	
4	01.53	02.83	03.70	4	
5	01.91	03.54	04.62	5	
6	02.30	04.24	05.54	6	
7	02.68	04.95	06.47	7	
8	03.06	05.66	07.39	8	
9	03.44	06.36	08.32	9	
10	03.83	07.07	09.24	10	
11	04.21	07.78	10.16	11	
12	04.60	08.49	11.09	12	
13	04.98	09.19	12.01	13	
14	05.36	09.90	12.93	14	
15	05.74	10.61	13.86	15	
16	06.12	11.31	14.78	16	
17	06.51	12.02	15.71	17	
18	06.89	12.73	16.63	18	
19	07.27	13.43	17.55	19	
20	07.65	14.14	18.48	20	
21	08.04	14.85	19.40	21	
22	08.42	15.56	20.33	22	
23	08.80	16.26	21.25	23	
24	09.18	16.97	22.17	24	
25	09.57	17.68	23.10	25	
26	09.95	18.38	24.02	26	
27	10.33	19.09	24.95	27	
28	10.72	19.80	25.87	28	
29	11.10	20.51	26.79	29	
30	11.48	21.21	27.72	30	
31	11.86	21.92	28.64	31	
32	12.25	22.63	29.56	32	
33	12.63	23.33	30.49	33	
34	13.01	24.04	31.41	34	
35	13.39	24.75	32.34	35	
36	13.78	25.46	33.26	36	
37	14.16	26.16	34.18	37	
38	14.54	26.87	35.11	38	
39	14.92	27.58	36.03	39	
40	15.31	28.28	36.96	40	
41	15.69	28.99	37.88	41	
42	16.07	29.70	38.80	42	
43	16.46	30.41	39.73	43	
44	16.84	31.11	40.65	44	
45	17.22	31.82	41.58	45	
46	17.60	32.53	42.50	46	
47	17.99	33.23	43.42	47	
48	18.37	33.94	44.35	48	
49	18.75	34.65	45.27	49	
50	19.14	35.36	46.20	50	
51	19.52	36.06	47.12	51	
52	19.90	36.77	48.04	52	
53	20.28	37.48	48.97	53	
54	20.67	38.18	49.89	54	
55	21.05	38.89	50.81	55	
56	21.43	39.60	51.74	56	
57	21.81	40.30	52.66	57	
58	22.20	41.01	53.59	58	
59	22.58	41.72	54.51	59	
60	22.96	42.43	55.43	60	
61	23.34	43.13	56.36	61	
62	23.73	43.84	57.28	62	
63	24.11	44.55	58.21	63	
64	24.49	45.25	59.13	64	
65	24.88	45.96	60.05	65	
66	25.26	46.67	60.98	66	
67	25.64	47.38	61.90	67	
68	26.02	48.08	62.83	68	
69	26.41	48.79	63.75	69	
70	26.79	49.50	64.67	70	
71	27.17	50.20	65.60	71	
72	27.55	50.91	66.52	72	
73	27.94	51.62	67.44	73	
74	28.32	52.33	68.37	74	
75	28.70	53.03	69.29	75	
76	29.09	53.74	70.22	76	
77	29.47	54.45	71.14	77	
78	29.85	55.15	72.06	78	
79	30.23	55.86	72.99	79	
80	30.62	56.57	73.91	80	
81	31.00	57.28	74.84	81	
82	31.38	57.98	75.76	82	
83	31.76	58.69	76.68	83	
84	32.15	59.40	77.61	84	
85	32.53	60.10	78.53	85	
86	32.91	60.81	79.46	86	
87	33.29	61.52	80.38	87	
88	33.68	52.22	81.30	88	
89	34.06	62.93	82.23	89	
90	34.44	63.64	83.15	90	
91	34.83	64.35	84.07	91	
92	35.21	65.05	85.00	92	
93	35.59	65.76	85.92	93	
94	35.97	66.47	86.85	94	
95	36.36	67.17	87.77	95	
96	36.74	67.88	88.69	96	
97	37.12	68.59	89.62	97	
98	37.50	69.30	90.54	98	
99	37.89	70.00	91.47	99	

TABLE IX

F=7.5MC

B_N	N		
	1 3 *5 *7 9 11 *13 *15 17 19 *21 *23 25 27 *29 *31	*2 6 *10 14 *18 22 *26 30	FOR C_{SIN}
	*1 3 5 *7 *9 11 13 *15 *17 19 21 *23 *25 27 29 *31	0 *4 8 *12 16 *20 24 *28	FOR C_{COS}
1	00.71	1	
2	01.41	2	
3	02.12	3	
4	02.83	4	
5	03.54	5	
6	04.24	6	
7	04.95	7	
8	05.66	8	
9	06.36	9	
10	07.07	10	
11	07.78	11	
12	08.49	12	
13	09.19	13	
14	09.90	14	
15	10.61	15	
16	11.31	16	
17	12.02	17	
18	12.73	18	
19	13.43	19	
20	14.14	20	
21	14.85	21	
22	15.56	22	
23	16.26	23	
24	16.97	24	
25	17.68	25	
26	18.38	26	
27	19.09	27	
28	19.80	28	
29	20.51	29	
30	21.21	30	
31	21.92	31	
32	22.63	32	
33	23.33	33	
34	24.04	34	
35	24.75	35	
36	25.46	36	
37	26.16	37	
38	26.87	38	
39	27.58	39	
40	28.28	40	
41	28.99	41	
42	29.70	42	
43	30.41	43	
44	31.11	44	
45	31.82	45	
46	32.53	46	
47	33.23	47	
48	33.94	48	
49	34.65	49	
50	35.36	50	
51	36.06	51	
52	36.77	52	
53	37.48	53	
54	38.18	54	
55	38.89	55	
56	39.60	56	
57	40.30	57	
58	41.01	58	
59	41.72	59	
60	42.43	60	
61	43.13	61	
62	43.84	62	
63	44.55	63	
64	45.25	64	
65	45.96	65	
66	46.67	66	
67	47.38	67	
68	48.08	68	
69	48.79	69	
70	49.50	70	
71	50.20	71	
72	50.91	72	
73	51.62	73	
74	52.33	74	
75	53.03	75	
76	53.74	76	
77	54.45	77	
78	55.15	78	
79	55.86	79	
80	56.57	80	
81	57.28	81	
82	57.98	82	
83	58.69	83	
84	59.40	84	
85	60.10	85	
86	60.81	86	
87	61.52	87	
88	62.22	88	
89	62.93	89	
90	63.64	90	
91	64.35	91	
92	65.05	92	
93	65.76	93	
94	66.47	94	
95	67.17	95	
96	67.88	96	
97	68.59	97	
98	69.30	98	
99	70.00	99	

TABLE X

F=8.75MC

B_N	N				
	1 7 *9 *15 17 23 *25 *31	*2 *6 10 14 *18 *22 26 30	3 5 *11 *13 19 21 *27 *29	*4 12 *20 28	FOR C_{SIN}
	*3 5 11 *13 *19 21 27 *29	2 *6 *10 14 18 *22 *26 30	*1 7 9 *15 *17 23 25 *31	0 *8 16 *24	FOR C_{COS}
1	00.38	00.71	00.92	1	
2	00.77	01.41	01.85	2	
3	01.15	02.12	02.77	3	
4	01.53	02.83	03.70	4	
5	01.91	03.54	04.62	5	
6	02.30	04.24	05.54	6	
7	02.68	04.95	06.47	7	
8	03.06	05.66	07.39	8	
9	03.44	06.36	08.32	9	
10	03.83	07.07	09.24	10	
11	04.21	07.78	10.16	11	
12	04.60	08.49	11.09	12	
13	04.98	09.19	12.01	13	
14	05.36	09.90	12.93	14	
15	05.74	10.61	13.86	15	
16	06.12	11.31	14.78	16	
17	06.51	12.02	15.71	17	
18	06.89	12.73	16.63	18	
19	07.27	13.43	17.55	19	
20	07.65	14.14	18.48	20	
21	08.04	14.85	19.40	21	
22	08.42	15.56	20.33	22	
23	08.80	16.26	21.25	23	
24	09.18	16.97	22.17	24	
25	09.57	17.68	23.10	25	
26	09.95	18.38	24.02	26	
27	10.33	19.09	24.95	27	
28	10.72	19.80	25.87	28	
29	11.10	20.51	26.79	29	
30	11.48	21.21	27.72	30	
31	11.86	21.92	28.64	31	
32	12.25	22.63	29.56	32	
33	12.63	23.33	30.49	33	
34	13.01	24.04	31.41	34	
35	13.39	24.75	32.34	35	
36	13.78	25.46	33.26	36	
37	14.16	26.16	34.18	37	
38	14.54	26.87	35.11	38	
39	14.92	27.58	36.03	39	
40	15.31	28.28	36.96	40	
41	15.69	28.99	37.88	41	
42	16.07	29.70	38.80	42	
43	16.46	30.41	39.73	43	
44	16.84	31.11	40.65	44	
45	17.22	31.82	41.58	45	
46	17.60	32.53	42.50	46	
47	17.99	33.23	43.42	47	
48	18.37	33.94	44.35	48	
49	18.75	34.65	45.27	49	
50	19.14	35.36	46.20	50	
51	19.52	36.06	47.12	51	
52	19.90	36.77	48.04	52	
53	20.28	37.48	48.97	53	
54	20.67	38.18	49.89	54	
55	21.05	38.89	50.81	55	
56	21.43	39.60	51.74	56	
57	21.81	40.30	52.66	57	
58	22.20	41.01	53.59	58	
59	22.58	41.72	54.51	59	
60	22.96	42.43	55.43	60	
61	23.34	43.13	56.36	61	
62	23.73	43.84	57.28	62	
63	24.11	44.55	58.21	63	
64	24.49	45.25	59.13	64	
65	24.88	45.96	60.05	65	
66	25.26	46.67	60.98	66	
67	25.64	47.38	61.90	67	
68	26.02	48.08	62.83	68	
69	26.41	48.79	63.75	69	
70	26.79	49.50	64.67	70	
71	27.17	50.20	65.60	71	
72	27.55	50.91	66.52	72	
73	27.94	51.62	67.44	73	
74	28.32	52.33	68.37	74	
75	28.70	53.03	69.29	75	
76	29.09	53.74	70.22	76	
77	29.47	54.45	71.14	77	
78	29.85	55.15	72.06	78	
79	30.23	55.86	72.99	79	
80	30.62	56.57	73.91	80	
81	31.00	57.28	74.84	81	
82	31.38	57.98	75.76	82	
83	31.76	58.69	76.68	83	
84	32.15	59.40	77.61	84	
85	32.53	60.10	78.53	85	
86	32.91	60.81	79.46	86	
87	33.29	61.52	80.38	87	
88	33.68	62.22	81.30	88	
89	34.06	62.93	82.23	89	
90	34.44	63.64	83.15	90	
91	34.83	64.35	84.07	91	
92	35.21	65.05	85.00	92	
93	35.59	65.76	85.92	93	
94	35.97	66.47	86.85	94	
95	36.36	67.17	87.77	95	
96	36.74	67.88	88.69	96	
97	37.12	58.59	89.62	97	
98	37.50	69.30	90.54	98	
99	37.89	70.00	91.47	99	

TABLE XI

TABLE XII

$C_{sin}=0$	$F=10$ Mc

For C_{cos}

Add the values of B_n for the following subscripts (retaining proper signs)

0 2 4 6 8 10 12 14 16 18 20 22 24 26 28 30

and add the values for the following subscripts (reversing their signs)

1 3 5 7 9 11 13 15 17 19 21 23 25 27 29 31

Appendix: Substitution of Sample Point Amplitudes A by Their Differences B in (7)

To solve this problem we introduce the following artifice: let the sequence $A_0, A_1, A_2, A_3, \cdots$ decrease by an attenuation factor $e^{-n\tau/\delta}$, so that the sequence becomes: $A_0, A_1e^{-\tau/\delta}, A_2e^{-2\tau/\delta}, \cdots$. For very large values of δ this will not affect our transient except for very large values of τ or—in the frequency domain—for very small values of ω.

If we use the relationship,

$$\sum_{m=0}^{n} B_m = A_n \tag{15}$$

(7) takes the form:

$$\Phi(\omega) = \frac{\pi}{\omega_c}\sum_{n=0}^{\infty}\left(\sum_{m=0}^{n} B_m\right)e^{-jn\tau\omega-n\tau/\delta}. \tag{16}$$

Writing $n\epsilon$ for $(jn\tau\omega+n\tau/\delta)$ it will be after several transformations,

$$\frac{\omega_c}{\pi}\Phi(\omega) = \left(\sum_0^{\infty} e^{-n\epsilon}\right)\left(\sum_0^{\infty} B_n e^{-n\epsilon}\right). \tag{17}$$

The series $\sum e^{-n\epsilon}$ is absolutely convergent if

$$\sum_0^{\infty}\left|e^{-n\epsilon}\right| = \sum_0^{\infty}\left|e^{-jn\tau\omega-n\tau/\delta}\right| \text{ converges.}$$

Since

$$\left|e^{-jn\tau\omega-n\tau/\delta}\right| = e^{-n\tau/\delta}$$

the series

$$\sum_0^{\infty}\left|e^{-j\omega n\tau-n\tau/\delta}\right|$$

can be considered to be a geometrical progression, which is known to be convergent for $e^{-\tau/\delta}<1$, i.e., the series is convergent for any positive, finite values of δ.

Using the formula for the sum of an infinite geometric progression, we obtain:

$$\sum_0^{\infty} e^{-n\epsilon} = \frac{1}{1-e^{-\epsilon}} = \frac{1}{1-e^{-j\omega\tau-\tau/\delta}}$$

$$= \frac{1}{1-e^{-\tau/\delta}\cdot e^{-j\omega\tau}}. \tag{18}$$

Since δ can be made arbitrarily large, the factor $e^{-\tau/\delta}$ can be made as close to unity as desired; we will, therefore, write:

$$\sum_0^{\infty} e^{-n\epsilon} = \frac{1}{1-e^{-j\omega\tau}} \tag{19}$$

or after a transformation:

$$\sum_0^{\infty} e^{-n\epsilon} = \frac{e^{j\omega\tau/2}}{2j\sin\omega\tau/2}. \tag{20}$$

Finally, we obtain (neglecting the constant time delay $e^{+j\omega\tau/2}$):

$$\Phi(\omega) = \frac{\pi}{\omega_c}\,\frac{1}{2j\sin\omega\tau/2}\sum_0^{\infty} B_n e^{-jn\tau\omega}. \tag{21}$$

With $\tau=1/2f_c=\pi/\omega_c$ we obtain (8).

Vertical Incidence Ionosphere Absorption at 150 Kc*

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Summary—The results of a year's experimental observations of the ionospheric absorption at vertical incidence at 150 kc are presented. In particular, the diurnal and seasonal variations of the absorption are examined. The existence of relationship between the variation of the absorption and the sun's zenith angle and vertical incidence critical frequency is established.

Introduction

THE ABSORPTION of radio waves in the ionosphere has been investigated by a number of workers, primarily at frequencies above the standard broadcast band. At the low end of the spectrum, scientists in England have explored the very long waves from 16 to 30 kc. That portion of the spectrum between 30 and 500 kc has, until recently, suffered a considerable lack of attention. This region is now becoming a center of active interest, partially because of its possible use for long-range navigational aids.

A rather cursory estimation of the daytime reflection

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coefficient as a function of frequency in this portion of the spectrum has indicated that this coefficient is near unity at the very low frequencies near 16 kc drops to a low minimum near 100 kc, and rises slowly again near 500 kc. The Radio Propagation Laboratory has been conducting measurements of virtual height and absorption at vertical incidence at 150 kc since February, 1949. It is the purpose of this paper to present a résumé of the experimental results obtained from the absorption measurements.

Equipment

Measurements were conducted on transmissions from a high-power pulse transmitter at a carrier frequency of 150 kc, a pulse rate of 1.5625 pps, and a Gaussian pulse shape 150 microseconds in width. The transmitter consists of a self-excited oscillator using ten 304TH triodes, pulsed by a hydrogen thyratron modulator through a pulse-forming network. The transmitting antenna is a folded dipole 3,000 feet long and 96 feet high. The receiving site is separated approximately 4 km from the transmitter, and an automatic record of the virtual height, the amplitude of the echo, and the amplitude of the ground pulse is made.

Investigations of the polarization of the echo indicated that under normal conditions it was elliptical, left-handed, and rather constant in orientation and configuration with variation of time; at least during night hours. This alleviated the necessity for attempting to devise a discriminator for use with the receiver in separating the two, non-split, magneto-ionic components. The receivers are superheterodyne type especially designed for this application. The height is recorded continuously and simultaneously from an oscilloscopic presentation by two cameras, one on a 150-km range and one on a 300-km range. The short-range record is scaled every fifteen minutes to an estimated accuracy of ± 1 km. The absorption is recorded by a servo on the gain of another receiver.

Experimental Procedure

The method of recording the reflection coefficient is the familiar procedure of Appleton.[1] The reflection coefficient ρ is given by:

$$\rho = \frac{2hE'}{k_1 G},$$

where ρ is the reflection coefficient, $2h$ is the total path, G is the amplitude of the ground pulse, E' is the amplitude of the first E-layer echo, and k_1 is a constant to be determined. k_1 may be determined by observation of the second E-layer reflection, E'', since

$$\rho^2 = \frac{E''4h}{k_1 G k_2},$$

[1] E. V. Appleton, "Regularities and irregularities in the ionosphere," *Proc. Roy. Soc.*, vol. 162, pp. 451–499; October, 1937.

where k_2 is the ground reflection coefficient. From Terman,[2] for a dielectric constant of 10 and a conductivity of 6×10^{-14} emu at 150 kc k_2 for both vertical and horizontally polarized waves at normal incidence is in order of 0.96. Therefore,

$$\rho = \frac{2E''}{0.96E'} = \frac{2hE'}{k_1 G}. \tag{1}$$

The constant k_1 may thus be determined by plotting hE'/G versus E''/E', and observing the slope of the least-square radial line passing through the experimental points. Such a calibration is performed at least once a week.

The reflection coefficient is related to the absorption coefficient k by $\rho = \exp(-\int k ds)$ where ds is the differential of the total path traversed by the wave and $\int k ds$ is the total absorption. Hence, the absolute value of the total absorption in nepers is equal to $|-\log_e \rho|$.

The records are scaled for G and E', and the reflection coefficient is computed from (1). $|\log_e \rho|$ is then plotted and a smooth mean curve drawn as shown in Fig. 1. The original records are examined for fine detail correlations such as sudden ionospheric disturbances, magnetic storms, and ionospheric storminess reported at short waves. The diurnal curves of $|\log_e \rho|$ are studied for relationship to the E-layer critical frequencies, the sun's zenith angle, and for seasonal trends.

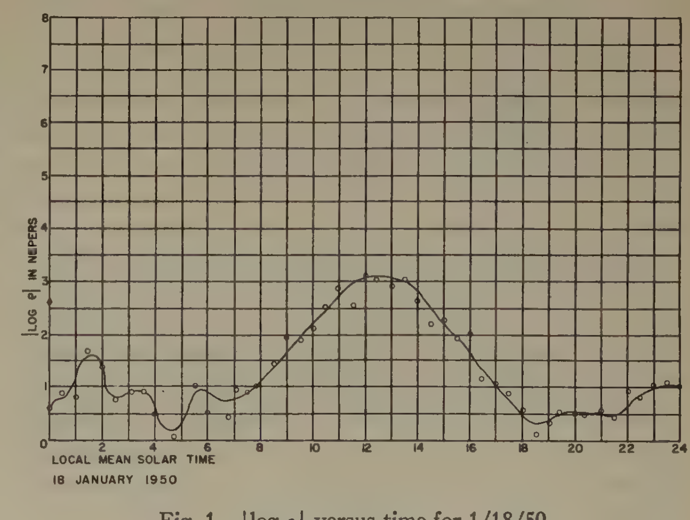

Fig. 1—$|\log \rho|$ versus time for 1/18/50.

Experimental Results

Since the virtual height enters into the calculation of the total absorption, some mention must be made of its general behavior. The virtual height of the layer as studied at 150 kc is similar to that reported by Helliwell[3] from measurements at 100 kc. From midnight the

[2] F. E. Terman, "Radio Engineers Handbook," McGraw-Hill Book Co., New York, N. Y.; 1943.

[3] R. A. Helliwell, "Ionospheric virtual height measurements at 100 kilocycles," Proc. I.R.E., vol. 37, pp. 887–894; August, 1949.

virtual height rises from a median value near 95 km to about 98 km at sunrise. At sunrise it drops rapidly about 10 km, then reaches a minimum near 85 km just after local mean solar noon to rise slowly again toward the midnight value. Splitting of the echo is quite prevalent, and in such cases the first echo is recorded. Tentative polarization measurements have indicated that such splits are not magneto-ionic.

Fig. 2 is a dual plot presenting the seasonal variation of the average nighttime absorption (from 2200 to 0200) and the maximum value of absorption attained

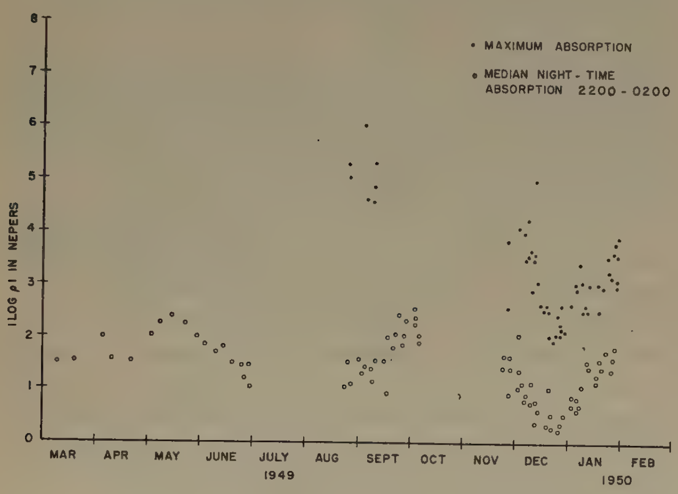

Fig. 2—Seasonal variation of maximum absorption and average nighttime value.

each day. As will be shown later, the maximum absorption does not normally fall at local noon. The gap in the maximum absorption during the summer months represents insufficient system gain to close the curve near noon. The gap between October and November is due to equipment modification. Evident from this figure is a minimum in absorption both in day and night near the winter solistic on December 21. During this time some short period nighttime reflection coefficients were greater than unity, which could be due to some phenomenon such as focusing. Although no noon-day records have been obtained near the summer solstice, the maximum absorption can be estimated to be in the order of 7 nepers. This value is obtained by extrapolating the $|\log_e \rho|$ curve from a known value of 0800 or 0900 to 1200 LMST by the relation

$$\left|\frac{\log \rho_1}{\log \rho_2}\right| = \left(\frac{\cos \chi_1}{\cos \chi_2}\right)^n \qquad (2)$$

where χ is the sun's zenith angle, and n is an exponent that will be determined below.

Appleton[4] has shown that the total absorption, theoretically, should follow $\cos \chi$ to the 3/2 power for nondeviating absorption such as is experienced by F-layer reflections. It is therefore logical to assume that the total absorption at these low frequencies should also follow some $\cos \chi$ law. It may be further assumed that the exponent will be smaller than 3/2, since in the low-frequency case we are dealing with deviating absorption in a level in the layer where collision is large. Computations have established that the quasi-longitudinal approximation to the Appleton-Hartree dispersion equation is valid for this application. Using this approximation,[5] a double parabolic fit to the Chapman electron distribution curve,[6] and an exponential collisional frequency versus height relation, a proportionality between the total absorption and $\cos \chi$ may be established.[7] Such a determination is dependent on the ability to accurately, experimentally fix the location of the height of the maximum ionization, and the value of the collisional frequency at some specified height, for a given value of the sun's zenith angle. These computations have given an exponent on $\cos \chi$ to be in the order of 0.8, but with an absolute value of $|\log_e \rho|$ that is too small. The data have been studied to determine this exponent. It is obtained by plotting $\log |\log_e \rho|$ versus $\log \cos \chi$. If the plot is linear, the exponent on $\cos \chi$ is the linear slope of the curve. An example of such a plot is shown in Fig. 3. The results of this study are presented in Table I.

The row entitled "total data" represents all days

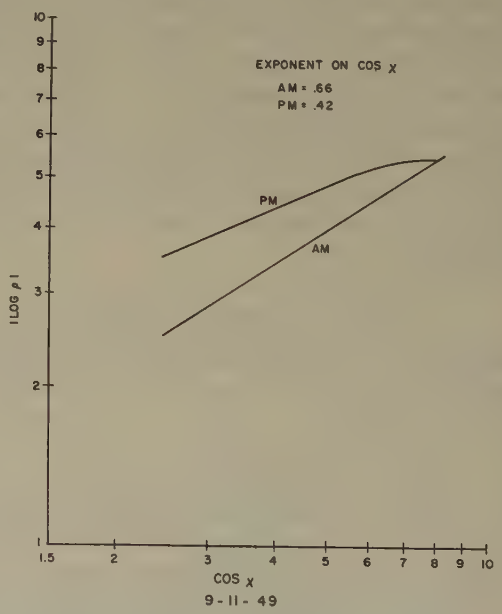

Fig. 3—$|\log \rho|$ versus $\cos \chi$ for 9/11/49.

[4] See page 458 of footnote reference 1.

[5] S. K. Mitra, "The upper atmosphere," *Roy. Asiatic Soc.* (Bengal), p. 168; Calcutta, India.

[6] J. E. Hacke, "An approach to the approximate solution of the ionosphere absorption problem," Proc. I.R.E., vol. 36, pp. 724–727; June, 1948.

[7] This problem will be examined in a later paper.

whose $|\log_e \rho|$ curves did not visually indicate any unusually perturbed condition. The second row and the third row represent the total data less any days in the CRPL-F bulletins[8] with a storm character number of 4 or greater on short-wave ionospheric storminess or geomagnetic storminess, respectively. No significant change is evident from this elimination. The fourth row gives the total data purged of days with evidences of sporadic E. Rather paradoxically, although the mean exponent has remained the same, the coefficient of variance has spread to a value of 54 per cent for the AM data. In the last row, all of the days with storms or sporadic E were cast out and in addition, only the very smoothest $|\log_e \rho|$ curves were considered. Both the morning and afternoon mean values dropped, and the PM coefficient of variance closed up to 37.6 per cent. The consistently smaller amount of data in the afternoon is a result of a lag in the ionization, which will be discussed later, causing the PM cases to be nonlinear. It should be emphasized that the object of this Table is to find the best value of X, and not to demonstrate the control of storms or sporadic E on these data. These values agree in order of magnitude with the value of the exponent cited above.

An important ionospheric quantity is the ratio of the noon-day values of the total absorption for summer and winter. Using the exponent 0.75 in (2) the ratio for the midday absorption between December 21 and June 21 is 1.8. This, however, does not check with the experimental value. The December 21 noon value is about 2 nepers, but the noon value at June 21 is much greater than 1.8 times this value. Using the estimated value of 7 nepers, the experimental ratio would be greater than 3.5. Such a seasonal anomaly has been reported by observers on nondeviating absorption at short waves. White and Straker[9] reported a value of 2.9 experimentally against 3.73 theoretically. Appleton[10] cites a value of 2.6 experimentally against a value of 6.4 theoretically for southeast England. Best and Ratcliffe[11] found that both the diurnal and seasonal changes obeyed the 1.5 exponent for selected days of F-layer reflection. It is of some significance that at 150 kc the exponent necessary to explain the seasonal change is considerably larger than that given by the diurnal curves, while the converse is true at short waves.

The 3/2 law on $\cos \chi$ for nondeviating absorption at short waves has been confirmed for selected days by Best and Ratcliffe.[11] A number of other investigators have measured this exponent and found it to be lower than the theoretical value of 3/2. These lower values have been attributed to the presence of an absorbing region below the normal E region, referred to as the D layer. It is hoped that more information can be obtained on this region from a thorough study of these low-frequency results.

Fig. 4 shows the relative coincidence of the point of "inflection" of the $|\log_e \rho|$ curve near sunrise and sunset to the variation of the actual ground sunset as given in the Ephemeris.[12] These curves show that there is a close correlation of the sunrise data, but that there is a noticeable lag in the sunset point of inflection. An explanation for this latter effect will be given below.

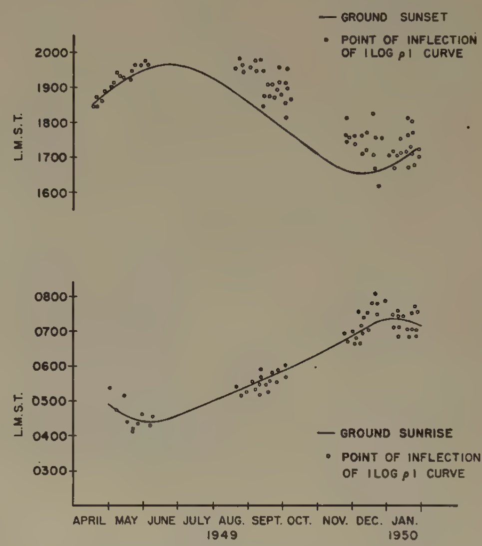

Fig. 4—Sunrise and sunset correlation.

An inspection of the complete diurnal curves of $|\log_e \rho|$ reveals that there is a distinct skewness or displacement of these curves past local noon, contrary to the simple Chapman theory. All available complete $|\log_e \rho|$ curves were examined for the point of maximum absorption, and a histogram of the results of this study is presented in Fig. 5. From this figure, it is evident that the maximum absorption occurs in the vicinity of 1300 hours LMST. This displacement or lag of the absorption with the sun's zenith angle corresponds with the lag noted in the sunset point of inflection in Fig. 4. Since we have seen that the AM value of $|\log_e \rho|$ varies as the 3/4 power of $\cos \chi$, and it has previously been established that the E-layer critical frequency varies as the 1/4 power of $\cos \chi$,[13] we then should expect $|\log_e \rho|$ to

[8] Ionospheric Data, CRPL-F, U. S. Dept. of Commerce, National Bureau of Standards, Central Radio Propagation Laboratory, Washington, D. C.

[9] F. W. G. White and T. W. Straker, "The diurnal variation of absorption of wireless waves," *Proc. Phys. Soc.*, vol. 151, pp. 865–875; September, 1939.

[10] See page 464 of footnote reference 1.

[11] J. E. Best and J. A. Ratcliffe, "The diurnal variation of the ionospheric absorption of wireless waves," *Proc. Phys. Soc.*, vol. 50, pp. 233–247; March, 1938.

[12] "The American Ephemerus and Nautical Almanac," United States Government Printing Office, Washington, D. C.; 1949.

[13] S. Chapman, "The absorption and dissociative or ionizing effect of monochromatic radiation in an atmosphere on a rotating earth," *Proc. Phys. Soc.*, Part I, vol. 43, pp. 26–45; January, 1931.

vary as the cube of the E-layer vertical incidence critical frequency f_c. Examining the records for such a relationship, we are immediately impressed by a similar skewness of these critical frequencies. Using critical frequencies taken from sweep-frequency records the AM value of exponent for 5 cases is 3.8 and the PM, 3.2. Using CRPL monthly means, the values are 2.85 for AM and 2.46 for PM. No other cases were examined because of the lack of high accuracy, local critical frequencies. For so few cases, these values check reasonably well with the value of 3.0. Because of a lag in the critical frequencies, the morning and afternoon values of these exponents are more nearly equal than are the exponents on cos χ.

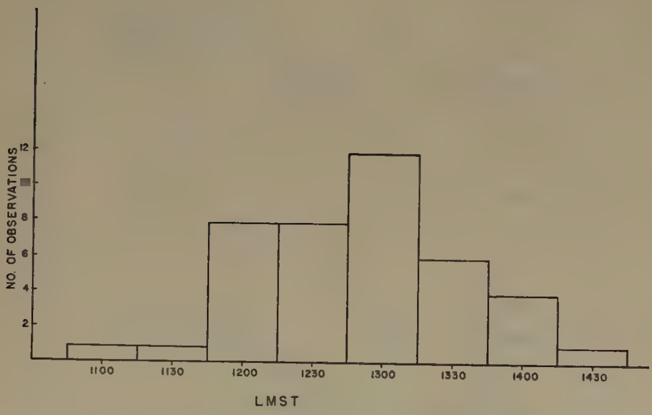

Fig. 5—Histogram of the location of the daily maximum absorption.

The lag in the diurnal variation of the ionization by observation of critical frequencies has previously been reported by Best, Farmer, and Ratcliffe,[14] and Kirby, Gilliland, and Judson.[15] This phenomenon can be explained in terms of recombination in Chapman's theory. Wilkes[16] has calculated the value of recombination coefficient from the above investigator's experimental results. He has shown that the lag is small where the layer is the densest, but that the lag increases at lower levels, which conforms nicely with the 150-kc results.

Conclusions

The analysis of the absorption data has brought out several important facts. First, the vertical incidence absorption at night is in the order of 1 neper, and the maximum absorption in the daytime varies from 2 nepers in midwinter to about 7 nepers in midsummer. Second, the exponent on cos χ that relates it to the total absorption is about 0.7 for the morning and 0.6 for the afternoon. Third, a seasonal anomaly exists on long waves. Finally, a lag in the absorption with the sun's zenith angle is quite predominant.

Acknowledgment

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TABLE I
Statistical Analysis on Exponent on Cos χ

	AM				*PM*			
	N	$\overline{X}$	σ	CV	N	$\overline{X}$	σ	CV
1. *Total Data*	41	0.75	0.265	35%	27	0.65	0.454	70%
2. *Without Ionospheric Storms*	39	0.763	0.26	35%	27	0.65	0.454	70%
3. *Without Magnetic Storms*	37	0.75	0.26	34.6%	27	0.65	0.454	70%
4. *Less Sporadic*	32	0.755	0.41	54%	25	0.654	0.46	71%
5. *Best Data*	10	0.62	0.202	32.5%	8	0.42	0.158	37.6%

Legend: N = the number of days
$\overline{X}$ = arithmetic mean of exponent on cos χ
σ = standard deviation
CV = coefficient of variation $= \sigma/\overline{X}$.

[14] J. E. Best, F. T. Farmer, and J. A. Ratcliffe, "Studies of region E of the ionosphere," *Proc. Roy. Soc.*, vol. 164, pp. 96–116; January, 1938.

[15] S. S. Kirby, T. R. Gilliland, and E. B. Judson, "Ionospheric studies during partial solar eclipse of February 3, 1945," Proc. I.R.E., vol. 24, pp. 1027–1041; July, 1936.

[16] M. V. Wilkes, "Theoretical ionization curves for the E region," *Proc. Phys. Soc.*, vol. 51, pp. 138–146; January, 1939.

Secondary-Emitting Surfaces in the Presence of Oxide-Coated Cathodes*

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Summary—**The experiments described here show that the deleterious effect of oxide cathodes on secondary-emitting surfaces of silver magnesium can be overcome by using tantalum instead of nickel as the base metal for the oxide coating.**

IT IS WELL KNOWN that secondary-emitting materials are subject to contamination from heated cathodes. The deterioration in secondary emission caused by this contamination is so serious that tubes using a hot cathode source and electron multiplication have usually[1–3] been specially designed so as to make the target ("dynode") less accessible to material coming from the cathode. According to a recent paper by Mueller,[4] this difficulty has been overcome for filamentary cathodes and also, to some extent, for indirectly heated cathodes.

The experiments described here were entirely confined to indirectly heated cathodes, and as they resulted in a novel means to avoid the contamination, a short report on them may be of interest.

I. Experimental Technique

A. Tubes

While an oxide cathode is being formed, its temperature is raised, for a short time, to a value which may be about 200° above its operating temperature. It is thus possible that the target contamination occurs, at least in part, during the forming process. Most of the experiments were therefore performed with tubes as shown in Fig. 1, with a sliding target which did not face the cathode during formation. The arrangement is cylindrical. The straight cathode K (diameter $\frac{1}{16}$ inch) occupies the center; it is surrounded by a collector which is shaped as a squirrel cage (24 tantalum wires, three of them 0.015 inch, the others 0.005 inch thick, held together by top and bottom Ta ribbons; outside diameter $\frac{3}{8}$ inch). The target is coaxial with the cathode and collector and can slide on three support rods S. On the pump the tube is mounted upside down, so that the target slides by gravity into the position shown in the drawing. In operation, it slides down to surround the collector.

* Decimal classification: R331. Original manuscript received by the Institute, May 11, 1950.

This work was done in 1948 under Contract W36-039-sc-33864 with the Evans Signal Laboratory.

† Capehart-Farnsworth Corporation, Fort Wayne, Ind.

[1] M. Chauvierre, "Secondary emission amplifier tube," *Tele-Tech.*, vol. 6, p. 69; July, 1947.

[2] H. M. Wagner and W. R. Ferris, "The orbital-beam secondary-emission multiplier for uhf amplification," Proc. I.R.E., vol. 29, pp. 598–602; November, 1941.

[3] C. S. Bull and A. H. Atherton, "A new secondary cathode," *Proc. IEE*, vol. 97, pp. 65–71; March, 1950.

[4] C. W. Mueller, "Receiving tubes employing secondary electron emitting surfaces exposed to evaporation from oxide cathodes," Proc. I.R.E., vol. 38, pp. 159–164; February, 1950.

A similar construction, without the movable features, was used on fixed-target tubes, in which the target is permanently exposed to the cathode.

Eight per cent, 4 per cent, and 1.7 per cent Mg-Ag alloy was used as target material, but the final experiments all used the 1.7 per cent alloy.

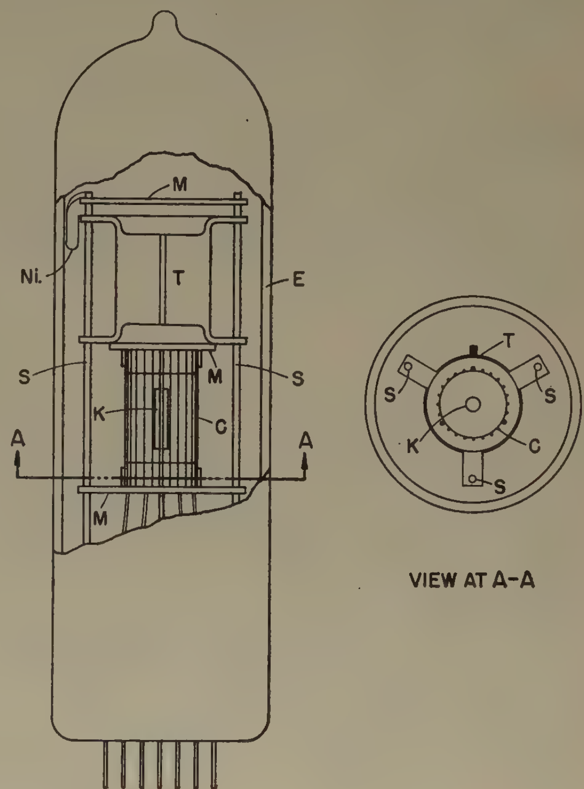

Fig. 1—Sliding target tube.
K = cathode
C = collector
T = target
M = mica disks
Ni = nickel ribbon
S = support rods
E = glass envelope.

B. Forming Schedules

The procedure which worked best was as follows: The tube, after assembly of parts, is given a vacuum bake-out at 460°C, and then the cathode temperature is raised in steps, so as to outgas it. The cathode is then flashed to about 1,150°C (true temperature), and then the collector is brought up to about 200 v, with the emission rising up to 80 ma. Thereafter, the cathode is cooled, oxygen is admitted, and the target is heated for about 1 minute with radio-frequency current. This oxidizes the target, but damages the cathode so that it

has to be re-formed. Then the getter (if a getter was used) is flashed and the target once more outgassed by radio-frequency current after which the tube is sealed off.

Numerous variations of this process (such as a bake in ozone) were used at one time or other.

C. Secondary Emission

The secondary emission was measured under standardized conditions: 100 volts on the target, 200 volts on the collector; the primary current density on the target was 2.6 ma per square centimeter.

The secondary emission coefficient K was defined as

$$K = \frac{I_c}{I_c - I_t} \tag{1}$$

where I_c and I_t are collector and target currents. This formula neglects the fraction α of the primary electrons which are intercepted by the collector without ever reaching the target. If this fraction is taken into account, it is found that the "true" secondary-emission coefficient K_t is connected to K as defined above by the formula

$$K - 1 = (K_t - 1)(1 - \alpha). \tag{2}$$

Attempts to determine α met only with a limited success; the best estimate is that $\alpha = 20$ per cent in our tubes. This would mean that $K = 3$ really corresponds to a "true" coefficient $K_t = 3.5$. The data recorded below refer to K rather than K_t, because K gives a more direct measure of the practical gain than can be realized in a secondary-emission tube.

The K measurements were performed with direct current, but in order to guard against the possible presence of time delays, as in the Malter effect,[5] it was ascertained that the collector current perfectly reproduces the cathode-emission variations, at least up to 1 Mc.

For life tests, it was necessary to keep the primary emission current of the cathode constant. This proved difficult, because any excess emission led to a noticeable increase in collector temperature which, by radiation, caused the cathode temperature to rise, thus increasing the emission still further. For the life tests, therefore, a thyratron circuit was used in the filament supply, controlled by the emission current so as to keep it constant.[6]

II. Results

Early results indicated that the decay in secondary emission depended very much on the cathode temperature. It became increasingly clear that the contamination came from the nickel base rather than from the oxide (barium-strontium carbonate, sprayed to a thickness of 0.002 inch with an amyl acetate binder, was used throughout). The base was a nickel sleeve (#799 DH in most cases). By a properly chosen formation schedule, we finally succeeded in achieving a constant secondary-emission ratio over several hundred hours on sliding targets. But even more constant results were obtained, on fixed as well as sliding targets, after the nickel sleeve had been replaced by tantalum. This was done on the assumption that a material which evaporated less easily than Ni should be used.

Fig. 2 shows life data taken on tube #365 (sliding target) and #368 (fixed target), both of which had a Ta sleeve as base for the oxide coating. As a further proof that Ta sleeves, even when run at overtemperature, cause no contamination, the record of tube #358

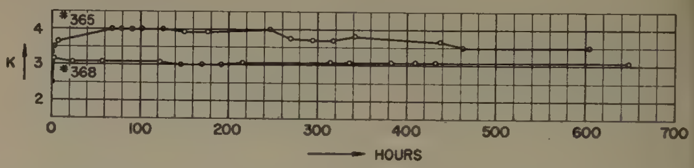

Fig. 2—Effective secondary-emission coefficient K of two representative tubes with tantalum cathode bases as a function of hours of operation.

is offered: This tube was operated at the usual emission current of 10 ma for 240 hours, showing a secondary-emission coefficient of $K = 3.2$. From 240 to 375 hours the emission was raised to 15 ma, giving $K = 3.4$. Thereafter, the emission was raised to 40 ma for 8 hours. During this time, values of K between 2.3 and 2.5 were measured, but it is almost certain that space charge formed between target and collector. When the emission was again lowered to 10 ma, K was 2.9 and stayed at this value. Similar, though less exacting, experiments with Ni always showed a rapid and permanent decay of K to values below 2.

One other experiment may be quoted to support the idea that the nickel sleeve is the source of contamination. In a tube of somewhat different construction it was possible to expose a target to a #799 Ni base without oxide coating. This base was heated to about 900°C. K dropped within 4 hours from its initial value of 3.2 to 1.9.

A similar experiment was performed in which Ba was evaporated. This, too, caused a rapid decay in secondary emission. The decay caused by an oxide cathode has commonly been ascribed to barium[3,4,7,8] while there are also accounts[9] which seem to minimize the importance of Ba. From our experiments, we are inclined to agree with this latter view.

[5] L. Malter, "Thin film field emission," *Phys. Rev.*, vol. 50, pp. 48–59; July 1, 1936.
[6] This circuit was designed and built by H. Beach.

[7] J. B. Johnson, "Secondary electron emission from targets of Ba-Sr oxide," *Phys. Rev.*, vol. 73, pp. 1058–1073; May 1, 1948.
[8] J. L. H. Jonker and A. J. Overbeck, "Application of secondary emission in amplifying values," *Wireless Eng.*, vol. 15, pp. 150–156; March, 1938.
[9] G. E. Moore and H. W. Allison, "Thermionic emission from thin films," *Phys. Rev.*, vol. 77, pp. 246–257; January 15, 1950.

III. Discussion

While these experiments prove that the main source of contamination comes from the Ni sleeve, there may be some doubt whether the nickel itself is to blame or some impurity in it. It may however be noted that Ni reaches a vapor pressure of 10^{-5} mm Hg at 1,160°C, i.e., slightly above the operating temperature, whereas Ta would have to be heated to 2,400°C in order to reach the same vapor pressure.[10] It is, therefore, quite possible that the contaminating substance is the nickel itself.

Using Ta, or some other metal with a low vapor pressure, instead of Ni, is entirely practical. It has been argued that oxide cathodes on a Ni base have a higher efficiency than those on other base metals. Our experiments have not extended in that direction; however, even if it should be found that the cathode temperature has to be slightly higher on Ta than on Ni, this would be a small price to pay for the resulting increase in tube life.

As good results were obtained in fixed-target tubes, it is seen that no special structure is required to prevent exposure of the target to the cathode during formation. This may be stated somewhat more generally as follows: Any contaminating agent which might come from the cathode during formation will be harmless if it is either not evaporated below the cathode-forming temperature (1,100°C) or else if it does not react with the target and can be re-evaporated below the target outgassing temperature (about 600°C).

[10] Data taken from S. Dushman, "Scientific Foundations of Vacuum Technique," John Wiley and Sons, Inc., New York, N. Y., and London, England, pp. 745–751; 1949.

On Poisoning of Oxide Cathodes by Atmospheric Sulfur*

H. A. STAHL†

By ELECTRON diffraction Huber and Wagener have found that on commercial oxide cathodes occasionally a face-centered space lattice appears.[1] According to the lattice constant ($d=6$, 37 A. u.), this lattice had to be attributed to barium sulfide. The question as to the origin, and effect of that sulfide on the emissivity was followed up during World War II.

This investigation was carried through both on colloidical mixed carbonates, after Buzágh and co-workers, and routine pastes precipitated from barium-strontium nitrate, or hydroxids. All the cathodes were either spread cataphoretically, or sprayed upon flat casings. In the latter case, an additional surface glazing by means of an amber roller was necessary. After conversion to oxide in the diffraction camera the cathodes were scrutinized at grazing incidence.

So as to identify the origin of cathodic sulfur, about 120 cathodes were distributed into several groups and stored for several months. They were laid on routine bakelite trays, on fluted cardboards, in Petri dishes, or were hung on nail boards, all exposed to the free access of air. Some control cathodes were sealed into glass bulbs, and evacuated.

These storage experiments showed that the strongest diffraction rings of BaS space lattice appear after a period fluctuating within wide limits from some days to several months, and that on all cathodes, except in sealed ones, nearly simultaneously. If so, the barium sulfide ring intensity gradually increased with storing period. As to the speed of sulfurizing, no reproducible results could be attained, however.

The observations suggested that cathodic sulfur originates from atmospheric air, the sulfur content of which varies, especially in cities and industrial districts, by a factor of 10^4. An average value is approximately 10^3 per cent by volume.

Tests on freshly made cathodes showed strong barium sulfide rings after exposing them for some seconds to hot combustion gases of a city gas-fed Bunsen burner. The same result appeared with cathodes exposed for 15 minutes to air containing 1 per cent sulfur dioxide, hydrogen sulfide, or carbon disulfide.

Emission measurements on sulfurized cathodes yielded a remarkable decrease of emissivity with increasing sulfur content (see Fig. 1). In accordance with Hiroshi Kamagawa's results on glasses rich in barium oxide,[2] it was concluded that on the cathode surface that was not broken down, a barium sulfite or sulfate was formed by atmospheric sulfur. This was thought to be reduced while converting at 1360° K, because of the carbon content of the collodian paste used. Attempts to confirm the existence of sulfite or sulfate on the spread carbonate surface failed, however.

It was felt worth while mentioning that several air-hygienists employ Ost's "baryta patch method," the main feature of which is barium carbonate as a detergent for atmospheric sulfur (both SO_2 and SO_3).[3] The action of barium carbonate in this case is, in several respects, much the same as on the cathode surface. Their manners of spreading differ by irrelevant subordinates only.

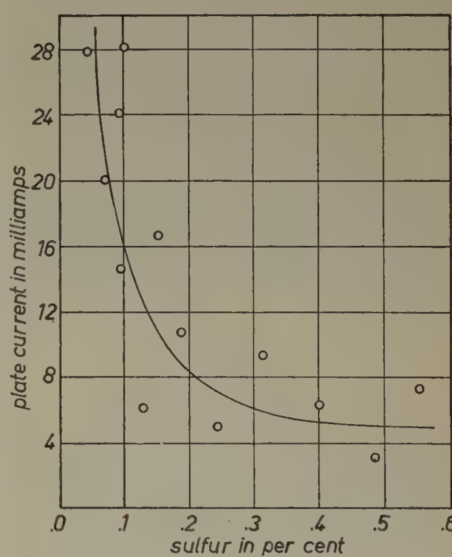

Fig. 1

* Decimal classification: R331. Original manuscript received by the Institute, March 21, 1949; abstract received, August 31, 1950. This paper has appeared in *Applied Scientific Research*, vol. B1, p. 397; 1950. Ir. C. W. Kosten, Editor, Mijnbouwplein 11, Delft, Holland.

† U. S. Troop Information and Education School, Stuttgart-Vaihingen, U. S. Zone, Germany.

[1] H. Huber and S. Wagener, "Die kristallographische Struktur von Erdalkalioxydgemischen," *Zeit. fur. tech. Phy.*, vol. 23, p. 10; 1942.

[2] Hiroshi Kamogawa, "Secondary emission and electron diffraction on the glass surface," *Phys. Rev.*, vol. 2, pp. 58, 660; 1940.

[3] H. Ost, "Schwefelsäure in der Atmosphäre," *Chem. Industrie*, vol. 23, p. 292; 1900. Also "Der Kampf gegen schadliche Industriegase," *Zeits. fuer angew. Chem.*, vol. 20, p. 1692; 1907. Also M. Bamberger and T. Nussbaum, "Luftuntersuchungen zur Feststellung Rauchschäden," *Z. angew. Chem.*, vol. 41, p. 23; 1928.

Correspondence

Amplification by Acceleration and Deceleration of a Single-Velocity Stream*

A method of amplification at microwave frequencies based upon the growth of space-charge waves in a decelerating stream of electrons has recently come to our notice. This mechanism became evident during a study of the type of waves described by Hahn[1] and Ramo.[2] It was found here that these waves not only change in length as the stream velocity changes, but also change in amplitude.

By a suitable combination of gradual decelerations and sudden accelerations, the amplitude of the space-charge wave may be essentially arbitrarily increased without the necessity of either wave carrying circuits, additional ions, or electron streams with different or distributed velocities, or space-charge-produced differences of velocity in a single stream.

An even simpler mechanism of amplification involving only short accelerating and decelerating gaps and constant potential drift regions exists which is closely related to the one just described. Consider a space-charge wave with ac velocity v_1 and an ac convection current density i_1 on a stream of electrons at a dc velocity u_1, described by

$$v = v_{1m} \cos\left(\frac{\omega_{p1}}{u_1} z\right) e^{j(\omega t - \omega z/u_1)} \qquad (1)$$

$$i_1 = j v_{1m} \frac{\omega}{\omega_{p1}} \frac{I_0}{u_1} \sin\left(\frac{\omega_{p1}}{u_1} z\right) e^{j(\omega t - \omega z/u_1)}, \qquad (2)$$

where $\omega_{p1}^2 = \eta I_0/\epsilon u_1$ and I_0 is the dc beam-current density. Now if at a position along the stream at which the ac velocity reaches its maximum value v_{1m}, the dc velocity is suddenly changed from u_1 to a lower value u_2, the ac velocity will increase from v_{1m} to v_{2m} such that

$$v_{2m} = v_{1m} \frac{u_1}{u_2}, \qquad (3)$$

provided only that the dc velocity change occurs in a distance which is short compared with a quarter space-charge wavelength at the lower velocity. That this is so can be demonstrated by simple kinematics or by application of the Llewellyn-Peterson diode equations.[3]

If the beam is then allowed to drift at the low velocity u_2 for an odd number of quarter space-charge wavelengths, that is, until the ac velocity has disappeared and the ac convection current which it produces is a maximum, this current i_2 will be

$$i_2 = i_{2m} = i_{1m}\left(\frac{u_1}{u_2}\right)^{3/2}, \qquad (4)$$

in which i_{1m} is the maximum ac convection-current density which would have been produced by the velocity modulation v_{1m}, if the stream had remained at the velocity u_1. At this point the stream may be suddenly returned to the dc velocity u_1. If this is again done in a distance which is short compared with the quarter space-charge wavelength at the lower velocity, the ac convection current is continuous across the gap, and the stream has returned to the original dc velocity with an ac current modulation which has been amplified from its original value by $(u_1/u_2)^{3/2}$. If the beam is again allowed to drift an odd number of quarter space-charge wavelengths, this current will convert to an ac velocity which is also $(u_1/u_2)^{3/2}$ times its original maximum value.

The dc velocity may be suddenly dropped again and the whole process repeated. Thus each stage consisting of one short low-velocity drift space and one long high-velocity drift space will provide an ac power amplification of $(V_1/V_2)^{3/2}$, where V_1 and V_2 are the dc voltages in the high- and the low-velocity drift spaces, respectively.

Amplification appears to be essentially independent of beam-current density, although the density determines the required lengths of the drift spaces, and the total beam current determines the maximum obtainable ac convection current, and hence the large signal saturation level.

An amplifier based on the above principles has been constructed and has provided a net power gain of 22 db at 3,000 Mc, using a single low-voltage drift region at 51 volts and two helices at 1,900 volts for modulation and demodulation of the stream. With the potential of the center drift region raised to 1,900 volts, the gain changed to zero db. Gain of progressively larger amounts was observed at drift region voltages of 178 volts, 117 volts, 78 volts, and 51 volts corresponding to n quarter space-charge wavelengths in the 5-cm drift space where n was 5, 7, 9, and 11, respectively. The total beam current was 0.7 ma and the approximate beam diameter, 0.15 cm.

At sufficiently low signal frequencies, the effective plasma frequency in the stream is reduced because of the finite beam size, and consequently the gain is reduced. At very high frequencies, it becomes difficult to excite the first-order plasma waves used in the above discussion, and higher-order space-charge waves will appear. It seems that they can be used to give gain, but require longer drift spaces and will saturate at lower power levels.

Finally, it might be mentioned that space-charge waves may be decreased as well as amplified by using similar principles, and where noise exists on the stream in the form of space-charge waves, the noise content in a limited frequency range may be reduced in this fashion. This has been experimentally verified in some low-noise traveling-wave amplifiers.

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* Received by the Institute, November 29, 1950. The research necessary for this correspondence was sponsored by the Office of Naval Research, the United States Army Signal Corps and the United States Air Force.

[1] W. C. Hahn, "Small signal theory of velocity modulated electron beams," *Gen. Elec. Rev.*, vol. 42, pp. 258–270; June, 1939.

[2] S. Ramo, "Space charge and field waves in an electron beam," *Phys. Rev.*, vol. 56, p. 276; August, 1939.

[3] F. B. Llewellyn and L. C. Peterson, "Vacuum-tube networks," PROC. I.R.E., vol. 32, pp. 144–166; March, 1944.

The Traveling-Wave Cathode-Ray Tube*

The paper by K. Owaki, S. Terahata, T. Hada, and T. Nakamura on "The Traveling-Wave Cathode-Ray Tube," in the October, 1950, issue of the PROCEEDINGS OF THE I.R.E. reveals significant progress in the field of microwave oscillography. In order to establish the development of this art—for 20 years one of the writer's hobbies—the following comments may be of some interest.

The prototype of the traveling-wave deflecting system are the multiphase deflecting plates.[1–3] According to Fig. 1(b) and (c), they consist of subsequent pairs of deflecting plates exhibiting alternate polarity due to their criss-cross connections. Maximum

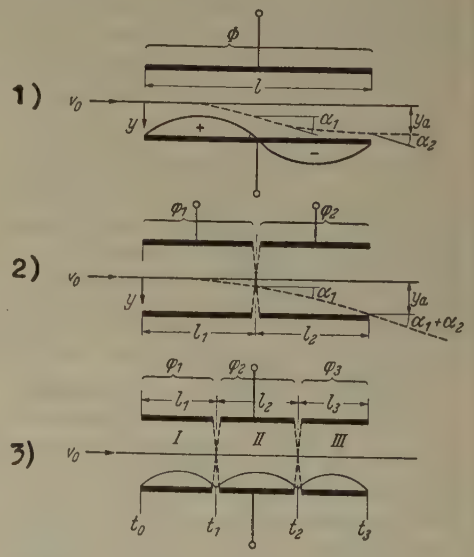


Fig. 1—(1) Single-, (2) two-, and (3) three-phase deflecting field.

* Received by the Institute, November 8, 1950.

[1] H. E. Hollmann, "Die Quersteuerung eines Kathodenstrahls in Mehrphasenfeldern," (Deflection of an electron beam in multiphase fields). *Elek. Nach. Tech.*, vol. 15, p. 336; 1938.

[2] H. E. Hollmann, "Das Verhalten der Kathodenstrahlröhre im Laufzeitgebiet," (The behavior of the cathode ray tube in the transit-time range), *Forts. der Hochfrequenz.*, vol. 1, p. 453; 1941.

[3] H. E. Hollmann, "Physik und Technik der Ultrakurzen Wellen," (Physics and technique of vhf), vol. 2, chapter 6, section 3, d; Berlin, 1936.

sensitivity occurs, of course, if frequency and beam velocity are matched in such a manner that the traveling electrons pass the partial fields always whenever they have the same polarity.

The improvement caused by the multiphase deflection, as compared with a single field (Fig. 1(a)) under dc operation, can be expressed by means of the multiphase inversion formulas:

$$P_1 = \frac{\sin\frac{\Phi}{2}}{\frac{\Phi}{2}} = \frac{1}{\Phi}\sqrt{2(1-\cos\Phi)}$$

$$P_2 = \frac{\sin^2\frac{\Phi}{4}}{\frac{\Phi}{4}} = \frac{2}{\Phi}\left(1-\cos\frac{\Phi}{2}\right)$$

$$P_3 = \frac{\sin\frac{\Phi}{2} - 2\sin\frac{\Phi}{6}}{\frac{\Phi}{2}} = P_1 - \frac{4}{\Phi}\sin\frac{\Phi}{6},$$

wherein Φ denotes the transit-time angle over the total deflecting system:

$$\Phi = \frac{\omega l}{v_0} = 2\pi\frac{lc}{\lambda v_0} = \frac{\pi l}{\lambda\sqrt{V_{p\,\text{volts}}}} \times 10^3.$$

(c = velocity of light; v_0 = beam velocity; V_p = plate voltage). The function P_1 is the almost classic inversion factor of a single field,[2-6] i.e., the dynamic sensitivity at any vhf referred to the static sensitivity. The functions P_2 and P_3 are the two- and three-phase versions. All three functions are diagrammed in Fig. 2. The two-phase system for dc produces no deflection whatsoever because the first partial field compensates the second field. The curve of the three-phase system starts at $\frac{1}{3}$ because only one partial field remains effective. The loss of static sensitivity, however, is compensated for by the shifting of the dynamic maxima towards higher Φ-values or higher frequencies, respectively. The first P_2-maximum occurs in the vicinity of 2π and P_3 in the vicinity of 3π which, in terms of present-day language, means accord between phase and beam velocity.

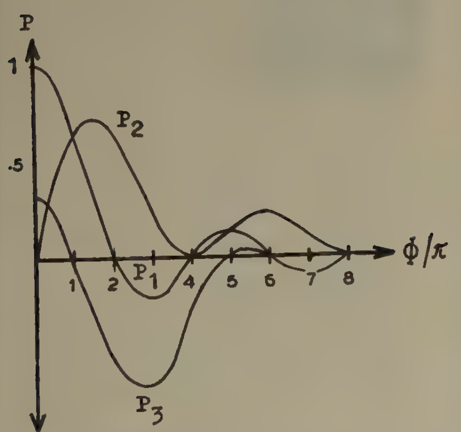


Fig. 2—The dynamic sensitivities of the three systems shown in Fig. 1 versus transit-time angle.

The inversion spectrograph[2,7,8] produces the inversion spectra shown in Fig. 3. The stray fields,[6] not included in the multiphase analysis, assure only a qualitative agreement between formulas and experiment. The fact that the maxima of an N-phase system remain below one and do not appear accurately at $N\pi$ is caused by the transit-time effects of the first kind, i.e., by the transit time elapsing in each individual field as well as by the phase-jumps.

Fig. 3—Experimental inversion spectra of the multiphase systems.

The disadvantage of the earlier multiphase systems with equal and adjacent fields can be overcome by various means. The simplest method is to diminish the axial length of the partial fields so that they operate quasi-statically with sufficient interspace in-between; however, this does not eliminate the stray field effects. Another method was applied by Pierce[9] in his multiphase or traveling-wave oscilloscope, wherein the former phase opposition is reduced by means of lumped-constant circuits, each feeding an individual pair of plates. From this device, only a short step leads to the traveling-wave oscilloscope described in Heaff's patent[10] and by the Japanese authors.

The ultradynamic Lissajous figures shown in the Japanese paper are the same as the writer's figures taken as far back as ten years ago.[2,8,11,12] The writer's method of a graphical analysis may well be applied to the Japanese figures. This may easily be understood because the traveling-wave system eliminates only the transit-time effect of the first kind but does not affect that of the second kind, namely, the transit time between both perpendicular deflecting fields.

All in all, the analogy between the step from the multiphase plates to the traveling-wave oscilloscope on the one hand and the step from the linear electron decelerator to the traveling-wave tube on the other hand is quite obvious.

HANS E. HOLLMANN
Oxnard, Calif.

[4] H. E. Hollmann, "Die Braunsche Röhre bei sehr hohen Frequenzen" (The cathode-ray tube at vhf), *Zeit. für Hochfrequenz.*, vol. 40, p. 97; 1932.
[5] H. E. Hollmann, "The use of the cathode-ray oscilloscope at ultra-high frequencies," *Wireless Eng.*, vol. 10, pp. 430 and 484; 1933.
[6] H. E. Hollmann, "The dynamic sensitivity and calibration of cathode-ray oscilloscopes at very high frequencies," PROC. I.R.E., vol. 38, p. 32; January, 1950.
[7] H. E. Hollmann, "Das Inversionsspektrum einer Braunschen Röhre," (The inversion spectrum of a cathode-ray tube), *Zeit. für. Tech. Phys.*, vol. 19, p. 259; 1938.
[8] H. E. Hollmann, "Ultra-high frequency oscillography," PROC. I.R.E., vol. 28, p. 213; 1940.
[9] J. R. Pierce, "Traveling-wave oscilloscope," *Electronics*, vol. 22, p. 97; November, 1949.
[10] A. Heaff, U. S. Patent No. 2,064, 469.
[11] H. E. Hollmann, "Ultradynamische Lissajous-Figuren," (Ultradynamic Lissajous figures), *Zeit. für Hochfrequenz.*, vol. 54, p. 19; 1939.
[12] H. E. Hollmann, "Mikrowellen-Oszillographie" (Microwave oscillography), *Zeit. für Hochfrequenz.*, vol. 54, p. 188; 1939.

Representation on Circuit Diagrams of Conductors in Contact*

In Fig. 14 of M. A. Schultz's paper[1] on "Linear Amplifiers," the lead to C26 and J4 (discriminator output) is drawn as making contact with the "ground" line. This is an obvious mistake, caused by putting a spot at the intersection of two conductors.

This occurrence prompts me to draw attention to the recommendation which has appeared for the last sixteen years in the British Standard Specification No. 530:[2]

"Of wires meeting at a connecting point, not more than two should be shown collinear."

They should be shown thus: 

and not: .

In my experience, neglect of this recommendation has been responsible for many mistakes, and I am careful never to draw a cross with a spot at the point of intersection.

I suggest that this recommendation might with advantage be adopted in an American Standard.

It is suggested that readers having comments on this subject address them to the chairman, IRE Symbols Committee, 1 East 79 Street, New York 21, N. Y.—*The Editor.*

L. H. BAINBRIDGE-BELL
Haslemere, Surrey
England

* Received by the Institute, June 22, 1950.
[1] M. A. Schultz, "Linear amplifiers," PROC. I.R.E., vol. 38, pp. 475–485; May, 1950.
[2] For further details, see L. Bainbridge-Bell, "Drawing circuit diagrams," *Wireless World*, vol. 55, pp. 179–180; May, 1949.

Contributors to the Proceedings of the I.R.E.

Arthur H. Benner (S'43–M'50) was born on March 29, 1922, in Leavenworth, Kan. He received the B.S. degree in electrical engineering from the University of Kansas in 1944, and the M.S. degree from the Pennsylvania State College in 1948. In 1950 he completed his work toward the Ph.D. degree at the Pennsylvania State College.

A. H. Benner

From 1946 to 1949 Dr. Benner was a research engineer with the Ordnance Research Laboratory, State College, Pa. From 1949 through 1950 he was associated with the Radio Propagation Laboratory of that College. He is a member of Sigma Xi and Tau Beta Pi.

❖

Robert C. Ellenwood (A'44) was born in Wooster, Ohio, on July 13, 1917. He received the B.S. degree from Ohio University in February, 1942, and after graduation he was employed by the Crosley Radio Corporation as a development engineer. During the war he served as a radio technician in the United States Navy. From 1945 until 1950 he was employed by the National Bureau of Standards as a research engineer to work on ultra-high-frequency standards and associated problems.

At present Mr. Ellenwood is an electronics engineer for the Electronics Shore Division, Bureau of Ships, Navy Department, in Washington, D. C.

❖

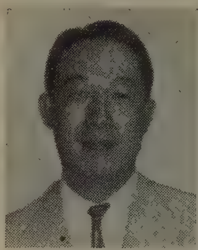
Gordon C. Dewey (A'47) was born in New York, N. Y., in 1923. He received the B.A. and the M.A. degrees in physics, both from Harvard University. Mr. Dewey was a wartime member of the staff of the Radiation Laboratory at the Massachusetts Institute of Technology. From February, 1947, to July, 1949, he was associated with the Federal Telecommunication Laboratories, Inc., during which time the work described elsewhere in this issue was carried out.

Gordon C. Dewey

Since September, 1949, Mr. Dewey has been a member of the Weapons Systems Evaluation Group which was formed in the National Military Establishment during February, 1949, under the sponsorship of the Joint Chiefs of Staff, the Office of the Secretary of Defense, and the Research and Development Board.

❖

Sze-Hou Chang (S'46–A'48) was born on September 23, 1913, in Ningpo, Chekiang, China. He received the B.S. degree in electrical communications from Chiao Tung University, Shanghai, in 1934. In 1946 he received the M.S. degree in communications engineering and, in 1948, the Ph.D. degree in engineering sciences and applied physics, both from Harvard University.



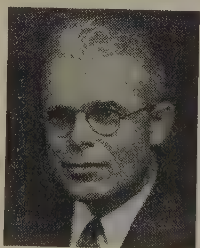
Sze-Hou Chang

During the period from 1934 to 1945, Dr. Chang served successively as assistant, lecturer, and professor in Tsing Hua University, Hunan University, and Chiao Tung University, all in China. He was associated with Cambridge Research Laboratories in Cambridge, Mass., as an electronic engineer from 1946 until 1948.

Dr. Chang is now an associate professor of electronic research at Northeastern University, Boston, Mass.

❖

Enoch B. Ferrell (A'25–M'29–SM'43) was born in Sedan, Kan., on June 1, 1898. He received the B.A., B.S. in electrical engineering, and M.A. degrees from the University of Oklahoma in 1920, 1921, and 1924, respectively. Mr. Ferrell taught in the department of mathematics of that university until 1924.



Enoch B. Ferrell

Since 1924 Mr. Ferrell has been a member of the research department of the Bell Telephone Laboratories, where he has been engaged in work on short-wave and ultra-short-wave radio transmitters, and on relays and switching systems for use in the telephone central-office plant.

❖

Martin W. Essigmann was born in Bethel, Vt., on January 14, 1917. He received the B.S.E.E. degree from Tufts College in 1938 and the S.M.E.E. degree from the Massachusetts Institute of Technology in 1947. He was appointed an instructor in electrical engineering at Northeastern University in 1938 and served in this capacity until 1944 when he became associated with the MIT Radar School. In 1947 Mr. Essigmann returned to Northeastern University where he is now an associate professor of electrical engineering.

M. W. Essigmann

Mr. Essigmann holds associate membership in the American Institute of Electrical Engineers and Sigma Xi, and he is a member of the American Association for the Advancement of Science, the American Society for Engineering Education, and Tau Beta Pi.

❖

Donald Glen Fink (A'35–SM'45–F'47) was born on November 8, 1911, in Englewood, N. J. He was graduated in 1933 from the Massachusetts Institute of Technology with the B.Sc. degree in electrical communications. After a year as a research assistant on the staff of the departments of geology and electrical engineering at MIT, Mr. Fink joined the staff of the journal *Electronics*, as editorial assistant. In 1937 he became managing editor; in 1945, executive editor; and in 1946, editor-in-chief. In 1942 he was awarded the degree of M.Sc. in electrical engineering by Columbia University.

Donald G. Fink

Obtaining a leave of absence from *Electronics* in 1941, Mr. Fink became a member of the staff of the Radiation Laboratory at MIT, where, in 1943, he headed the loran division. He then transferred to the Office of the Secretary of War as an expert consultant on radio navigation and radar. During his war service Mr. Fink traveled over 80,000 miles from Cairo, Egypt, to Darwin, Australia, siting loran stations and arranging for the use of the loran system by the allied forces. He participated in the atom bomb tests at Bikini, also.

Mr. Fink is the author of numerous books, including "Engineering Electronics," "Principles of Television Engineering," and "Radar Engineering." As editor of the

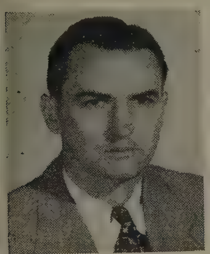
Contributors to the Proceedings of the I.R.E.

Proceedings of the National Television System Committee, member of the Television Panel of the Radio Technical Planning Board, and currently as a member of the Joint Technical Advisory Committee, he is active in standardization work, particularly in the field of television.

In 1948, Mr. Fink was chairman of the IRE Television System Committee, and in 1950 he was a member of the Senate Advisory Committee in Color Television. He is also a member of the Committee on Navigation of the Research and Development Board. He is a member of Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

❖

Theodore S. George was born on October 10, 1911, at Grove City, Pa. He received the B.S. degree in mathematics from Grove City College in 1932, and the M.A. and Ph.D. degrees in mathematics from Duke University in 1936 and 1942, respectively. From 1938 to 1945 he served as instructor and assistant professor of mathematics at the University of Florida. During the period from 1942 to 1945 he was on military leave as a Naval electronics officer, leaving the Navy with the rank of lieutenant commander. During this time, he served at sea as radar officer aboard a carrier and later in the Bureau of Aeronautics in charge of development of electronic bombing and fire-control devices.



T. S. George

Since the war Dr. George has been a consulting engineer in the research division of the Philco Corporation doing theoretical work in a variety of electronic problems.

❖

William C. Jakes, Jr. (S'43–A'49) was born in Milwaukee, Wis., on May 15, 1922. He received the B.S.E.E. degree from Northwestern University in 1944, then entered the Navy and served for two years as airborne electronics officer. He returned to Northwestern University in 1946 for graduate study, receiving the M.S. degree in 1947 and the Ph.D. degree in 1949. During part of this time he was employed by the Microwave Laboratory at Northwestern as a research associate.

W. C. Jakes, Jr.

Since July, 1949, Dr. Jakes has been a member of the technical staff of the Bell Telephone Laboratories, Inc., engaged in microwave propagation and antenna studies. He is a member of Sigma Xi, Eta Kappa Nu, and Pi Mu Epsilon.

❖

Theodore J. Marchese (A'44–SM'50) was born in Carlstadt, N. J., on October 17, 1912. In 1932 he was employed by the Federal Telegraph Company as a technician in their vacuum-tube department and he worked at a number of technical and supervisory jobs until 1941, when he joined the engineering department of Federal Telephone and Radio Corporation as a vacuum-tube engineer on large tubes.

T. J. Marchese

In 1947 he transferred to the Federal Telecommunication Laboratories, Inc., as an engineer on microwave power generators, and is presently employed in the development of negative-grid and traveling-wave amplifier tubes.

He received the B.S. degree in electrical engineering in 1948 from the evening school of Newark College of Engineering.

❖

Eleanor M. McElwee was born in New York, N. Y., in 1924. She received the A.B. degree in English and mathematics from Ladycliff College, Highland Falls, N. Y., in 1944, and has done additional work in science at Cooper Union.

Miss McElwee was employed by the Western Electric Tube Shop in New York, N. Y., from 1944 to 1947 as assistant product engineer. She joined the staff of Sylvania Electric Products Inc., in 1947, and was in charge of the statistical analysis program and life-testing of tubes for the Product Development Laboratories at Kew Gardens, L. I., N. Y., until early in 1950. She is at present a member of the editorial and information section at the same Laboratory.

Eleanor McElwee

Scott Nevin was born on September 11, 1924, in Ithaca, N. Y. He received the B.M.E. degree from Cornell University in 1945.

He joined the Capehart-Farnsworth Corporation in Fort Wayne, Ind., in 1946. Since that time, Mr. Nevin has worked on photocathodes. His recent research has also included the study of secondary-emission devices.

Scott Nevin

❖

Philip Parzen was born on June 28, 1916, in Poland. He received the B.S. degree in physics from the College of the City of New York in 1939 and the M.S. degree in physics from New York University in June, 1946. He is at present completing his work for the Ph.D. degree in mathematics at New York University.

During the war he was employed at the Westinghouse Research Laboratories and, since 1947, at the Federal Telecommunication Laboratories, Inc., working on problems in microwave tubes and electromagnetic wave propagation.

Philip Parzen

Mr. Parzen is a member of the American Physical Society.

❖

George E. Pihl (A'41) was born on February 22, 1915, at Brockton, Mass. He received the B.S. degree in electrical engineering from Northeastern University in 1937 and the M.S. degree in communications engineering from the Harvard Graduate School of Engineering in 1939.

Since 1938 Mr. Pihl has been a member of the faculty of Northeastern University, where he is at present an associate professor. He is a member of the American Institute of Electrical Engineers and Tau Beta Pi.

George E. Pihl

Contributors to the Proceedings of the I.R.E.

Hugo F. Pit was born in Utrecht, Netherlands, on April 3, 1924. Having attended the Haarlem Gymnasium, he is now studying electrical engineering at the Delft Institute of Technology.


Hugo F. Pit

The study of oscillator behavior in this issue of the Proceedings was a result of his activities in the Central Laboratory of the Netherlands Postal and Telecommunications Services during the summer of 1949.

❖

Evert J. Post was born in Rotterdam, Netherlands, on October 20, 1914. He received the degree of Physical Engineer from the Delft Institute of Technology.


Evert J. Post

In 1946 Mr. Post joined the Central Laboratory of the Netherlands Postal and Telecommunications Services at The Hague where he has been primarily concerned with the development of quartz crystals and oscillators.

❖

Mr. William E. Ryan (S'43–A'45) was born in Springfield, Mass., on February 6, 1920. He received the B.S. in both electrical engineering and mathematics from Michigan University in 1943. He was enrolled in evening classes at George Washington University Graduate School in Washington, D. C., from 1946 to 1950. From 1943 to 1944 he was employed by the Naval Research Laboratory as a radio engineer assisting in tests of antiaircraft gun directors.


William E. Ryan

Since 1944, Mr. Ryan has been employed by the National Bureau of Standards, assisting in the development of field intensity standards. At present he is associated with the Ultra-High Frequency Standards Group of the Microwave Standards Section of CRPL in Washington, D. C. He is a member of Eta Kappa Nu and Sigma Pi Sigma.

❖

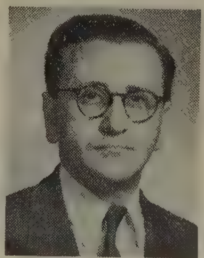
Hans Salinger (A'37–SM'46–F'51) was born in Berlin, Germany, on April 1, 1891. In 1915, he received the Ph.D. degree from the University of Berlin. From 1919 to 1929 he was a research associate at the Reichspostzentralamt and, from 1929 to 1935, a professor at the Heinrich Hertz Institute, both in Berlin. In 1936, he joined Farnsworth Television, Inc., and is now a research engineer with the Capehart-Farnsworth Corporation in Fort Wayne, Ind. He is also doing part-time teaching at the Purdue University Extension Center in Fort Wayne.


Hans Salinger

Dr. Salinger has published numerous papers on filters, measuring methods, electron multipliers, and other subjects. He is a member of Sigma Xi, the American Association for Advancement of Science, and the American Physical Society.

❖

H. A. Samulon (A'46) was born in Graudenz, Germany, in 1915. He attended the Swiss Federal Institute of Technology in Zurich, graduating with the degree of dipl. ing. in 1939. After doing postgraduate work at the Institute of High-Frequency Techniques, he was employed from 1943 to 1947 as an instructor and research engineer at the Institute of Communication Techniques and the Acoustical Laboratory of the Swiss Federal Institute of Technology.


Henry A. Samulon

In 1947, Mr. Samulon joined the Electronics Department (Electronics Laboratory Division) of the General Electric Company, where at present he is mainly engaged in TV problems, with special regard to transient phenomena as well as system problems in black-and-white and color television.

❖

Howard E. Sorrows (A'48) was born in Hewitt, Texas, on August 10, 1918. He received the B.A. degree in mathematics from Baylor University in 1940. After a year as instructor of mathematics and physics in the Sabinal, Texas, Public High School, he joined the technical staff of the National Bureau of Standards. He attended night classes of the George Washington Electrical Engineering School from 1941 to 1945, following which he entered their Graduate School, receiving the M.A. degree in physics in 1948. At present he is enrolled in the University of Maryland Graduate School. From 1941 to 1943 he assisted in various phases of the program for the development of field intensity and voltage standards. During World War II, he assisted in several projects of the radar-electronic countermeasures program conducted by the National Bureau of Standards for the Bureau of Ships. From 1947 to present he has been in charge of the Ultra-High Frequency Standards Group, Microwave Standards Section, CRPL, Washington, D. C.


H. E. Sorrows

Mr. Sorrows was a student member of AIEE and is an associate member of Sigma Pi Sigma and Sigma Xi. At present he is a member of the AIEE Sub-Committee Radiation Measurements above 300 Mc.

❖

Raymond E. Zenner (M'46–SM'50) was born in Chicago, Ill., on April 16, 1910. He received the B.S. degree in physics from the University of Chicago in 1933, and has pursued graduate studies at the Illinois Institute of Technology.


R. E. Zenner

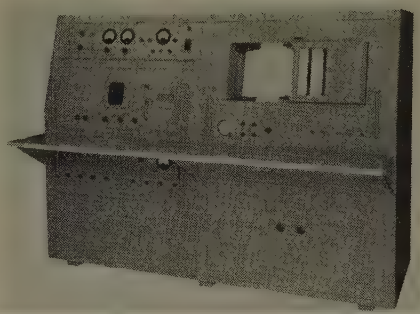
He has been employed in development and research work with the Teletype Corp., the A.B. Dick Co., Eicor, Inc., and the Armour Research Foundation, all of Chicago, on printing telegraph systems and terminal equipment, mimeograph products, magnetic recording research and product design, ordnance development, and tactical situation simulators for training military personnel. He is now supervisor of instrumentation and recording, electrical engineering department, at Armour Research Foundation.

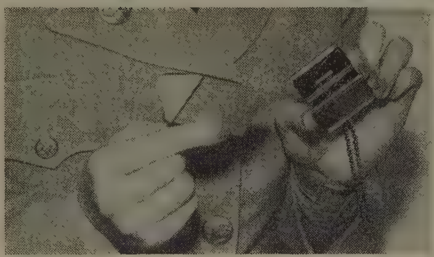
Mr. Zenner is serving on the Sound Recording and Reproducing Committee, the Magnetic Recording Subcommittee, and the Annual Review Committee of the IRE, and also on the RMA Magnetic Recording Subcommittee.

Institute News and Radio Notes

Technical Committee Notes

The Standards Committee, under the Chairmanship of Professor J. G. Brainerd, held a meeting on December 14. The following Standards have been approved by this Committee: **Proposed Standard on Electroacoustics: Definition of Terms, prepared by the Electroacoustics Committee of IRE;** and **Standard Abbreviations of Radio Electronic Terms,** prepared by the Symbols Committee. . . . On November 28, the **Wave Propagation Committee** held a meeting, at which C. R. Burrows presided as Acting Chairman. Reports on the activities of the various Subcommittees were given by the Chairmen. . . . The **Receivers Committee** under the Chairmanship of R. F. Shea held a meeting on November 1. Reports on the progress of work going on in the Subcommittees were given by the Chairmen. This Committee is working towards the completion of a **Standard on Radio Receivers: Methods of Measurement of Spurious Radiation, Frequency Modulation and Television Receivers.** It is expected that this Standard will be available in the Fall of 1951. . . . A meeting of the **Video Techniques Committee** was held on November 16, under the Chairmanship of J. E. Keister. Subcommittee 23.3, Video Systems and Components, prepared two tutorial papers which appeared in the January issue of the PROCEEDINGS, and work towards the preparation of more tutorial papers is continuing. The Video Techniques Committee also prepared a **Standard on Television: Methods of Measurement of Electronically Regulated Power Supplies,** published in the January issue of the PROCEEDINGS, reprints of which may now be purchased from Headquarters. . . . **Reprints of Standards on Circuits: Definitions of Terms in Network Topology,** may also be purchased at Headquarters. . . . **The proposed Standards on Electronic Computers: Definition of Terms, prepared by the Electronic Computers Committee of the IRE,** has been approved as an IRE Standard and will be published in the March issue of the PROCEEDINGS . . . Axel G. Jensen, Chairman of the Definitions Co-ordinating Subcommittee, has distributed to all Technical Committee Chairmen copies of a Proposed Standard on Definitions of Receiver Terms, prepared by the Receivers Committee, and also a proposed definition for the term "transfer characteristics." This is in line with the new procedure to obtain comments on the definitions by mail prior to submission to the Standards Committee for approval. . . . **A Joint IRE/AIEE Conference on Electron Tubes for Computers was held on** December 11 and 12 at the Haddon Hall Hotel, Atlantic City, N. J. Members of the IRE Technical Committee on **Electron Tubes and Solid-State Devices,** and the Committee on **Electronic Computers,** participated in formulating plans for this Conference. The sessions were well attended and enthusiastic.





TO BE DISPLAYED AT IRE CONVENTION

The devices shown above, which will be on display at the Armed Services exhibit, are typical of the wide variety of electronic equipment to be shown at Grand Central Palace during the IRE Convention.

At top is the console of a Cloud Base and Top Indicator, capable of measuring the thickness of overhead cloud decks from 300 to 50,000 feet, through rain if necessary. Continuous facsimile recordings are provided.

At bottom is a new device for measuring the extent of exposure of individuals to atomic radiation. The dosimeter, which is worn about the neck, gives readings one minute after exposure through a self-developing photographic process.

Calendar of COMING EVENTS

1951 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 19–22

URSI Spring Meeting, Washington, D. C., April 16–18

IRE Southwestern Conference, Dallas, Texas, April 20–21

1951 Convention of SMPTE, April 30–May 4, Hotel Statler, N. Y.

1951 Annual Meeting of the Engineering Institute of Canada, Mount Royal Hotel, Montreal, May 9–11

1951 IRE Technical Conference on Airborne Electronics, Biltmore Hotel, Dayton, Ohio, May 23–25

1951 IRE 7th Regional Conference, University of Washington, Seattle, Wash., June 20–22

1951 Summer General Meeting of AIEE, June 25–29, Royal York Hotel, Toronto, Canada

1951 IRE West Coast Convention, San Francisco, Calif., August 29–31

1951 IRE Convention Slated for March 19–22 in New York

The 1951 IRE National Convention, to be held on March 19–22 in New York, N. Y., promises to be the largest and most important convention in Institute history. An extensive technical program of over 200 paers will be presented in 43 sessions and symposia at the Waldorf-Astoria Hotel, Belmont Plaza Hotel, and Grand Central Palace. The Radio Engineering Show, comprising 269 exhibits of electronic and communications equipment and their applications, will fill three floors of Grand Central Palace.

The Convention will open with the Annual Meeting of the Institute on Monday morning, March 19, in the Grand Ballroom of the Waldorf. James W. McRae, director of transmission development for Bell Telephone Laboratories, will be the principal speaker. In addition, reports will be heard from IRE officers concerning the operations of the Institute during 1950. The program of technical sessions will commence on Monday afternoon.

A noteworthy feature of the technical program is the prominent part played by IRE Professional Groups in organizing symposia in their respective fields of interest.

Fourteen symposia are scheduled for this year's program, eleven of which are the result of Professional Group activities. The sponsoring Groups and the titles of their symposia are as follows: Audio Group—"Loudspeakers"; Broadcast Transmission Systems Group—"Broadcast Transmission Systems" and "Panel Discussion on the Empire State Story"; Circuit Theory Group—"New Extensions of Network Theory"; Instrumentation Group—"Amplification of DC Signals," "Panel Discussion on Performance of DC Amplifiers," and "Industrial Instrumentation"; Nuclear Science Group—"Nuclear Reactors"; Quality Control Group—"Panel Discussion on Tube Reliability"; Radio Telemetry and Remote Control Group—"Telemetering Systems" and "Simulation as an Aid To Design of Remote Control Systems."

Symposia will also be held on the following subjects: "Matching Schools and Industry" (sponsored by the IRE Education Committee), "Color Television," and "Some Systems Problems of Air Traffic Control."

The complete program, with abstracts of all papers, will be published in the March issue of the PROCEEDINGS.

The attention of broadcast engineers is called to "Broadcast Day" on Tuesday, March 20, when two Broadcast Group symposia (as noted above) will be held in the morning and afternoon, climaxed by the Color Television symposium on Tuesday night.

Of particular interest to all Professional Group members will be the President's Luncheon on Tuesday where tables will be assigned to each Group, affording Group members a unique opportunity to broaden their contacts within their own fields.

Tutorial Papers Series Begins in this Issue

The first paper in this issue, "Television Broadcasting in the United States, 1927–1950," (pp. 116–123) by Donald G. Fink, marks the beginning of a series of tutorial papers on a wide variety of subjects which are to be published at frequent intervals in the PROCEEDINGS OF THE I.R.E. during 1951 and 1952. These papers are to be educational in nature, of exceptional clarity, and prepared by leading authorities on topics of both present and historical interest.

These tutorial papers are procured by, and published on the recommendation of the Subcommittee on Tutorial Papers (Prof. Ernst Weber, Chairman) of the IRE Committee on Education (Prof. Herbert J. Reich, Chairman). It is believed that members in all grades will derive considerable benefit from this series of tutorial papers.

Sarnoff Gold Medal Award Is Established by SMPTE

The establishment of the David Sarnoff Gold Medal as an annual award for an outstanding contribution to television engineering has been announced by the Society of Motion Picture and Television Engineers.

The award will be presented at the Society's fall meeting to that individual who has done outstanding work in some technical phase of the broad field of television engineering, whether in research, development, design, manufacture or operation, or in any similar phase of theater television.

Terminology for IRE Publications

The publication activities of the Institute may be expanded at some date in the future not as yet determined. In order to avoid confusion, the membership is urged to use the following terminology in their discussions and correspondence when referring to existing and possible future IRE publications:

1. PROCEEDINGS OF THE I.R.E.—the official monthly publication of the Institute.

2. TRANSACTIONS — publications sponsored by IRE Professional Groups, either in conjunction with or independent of Professional Group symposia and conferences.

3. CONVENTION JOURNAL—a publication containing all papers presented at an IRE national convention.

Adoption of the above terminology in no way signifies that plans for the latter two publications are necessarily contemplated or as yet approved by the Institute.

First Call for Papers for West Coast IRE Convention

Authors are invited to submit prospective papers for the 1951 West Coast IRE Convention, particularly in the fields of Antennas, Circuits, Computers, Propagation, and Vacuum Tubes. The Convention will be held in San Francisco on August 29, 30, and 31. The following information should be submitted: (1) Name and address of author, (2) title of paper, and (3) a 100-word abstract and such additional information as may be required in order to properly evaluate the paper for inclusion in the technical program.

Please address all material to J. V. N. Granger, Stanford Research Institute, Stanford, Calif. The deadline for acceptance is May 1, 1951. Your prompt submission will insure full consideration.

Deadline Set for Papers for IRE 7th Regional Conference

Persons or companies are invited to submit prospective papers for the IRE 7th Regional Conference of 1951, to be held at the University of Washington, Seattle, Wash., on June 20, 21 and 22. In addition, manufacturers are invited to display equipment specifically tied in with the papers they may present. The following information should be submitted: (1) Name of paper, (2) author, (3) person or persons who will present it, (4) synopsis and (5) will equipment tieing in with the paper be displayed.

The deadline for acceptance is March 31. Please send all material to J. E. Hogg, Electronics Department, General Electric Co., 1146 Dexter Horton Building, Seattle, Wash.

NAB Conducts Third Annual Survey of TV Employment

Approximately 8,500 persons comprise the staffs of the TV stations and networks now on the air, according to an estimate announced by Richard P. Doherty, director of the National Association of Broadcasters' Employee-Employer Relations Department, in the third NAB annual TV employment survey. The compilation was based on information supplied by 56 TV stations (exclusive of networks) for a typical operative week during the late spring of 1950.

The increasing number of television stations which have become operative during the past year has reduced the "per station" employment average from 60 to 57 persons, according to the findings of the survey. The "per station" employment decline is a statistical consequence of the operational increase within the industry. The 1949 survey revealed the average station (exclusive of networks) employed 66 persons (46 full-time and 20 part-time); the mean in the spring of 1950 was 57 (39 full-time and 18 part-time). The reasons, in part, for this drop are that new stations are on the air less time than the established telecasters; and, for the most part, are located in smaller cities where over-all operation permits a smaller staff.

On the basis of the data collected in 1950, the average staff of 57 persons was made up of the following figures:

Average Staff per TV Station	*Number Full-Time Employees*	*Number Part-Time Employees*
Technical	18	1
Film	3	0
Program	10	10
Administrative	6	6
Sales	2	1
Total	39	18

IRE Official Appointed Selective Service Advisor

The six Scientific Advisory Committees, appointed in 1948 by Major General Lewis B. Hershey, Director of the Selective Service System, have recently submitted to the Director a joint report presenting recommendations concerning the training and utilization of scientific, professional, and specialized personnel. The committees represent the six following fields: Agricultural and Biological Sciences, Engineering Sciences, Healing Arts, Humanities, Physical Sciences, and Social Sciences. George W. Bailey, Executive Secretary of the IRE, was appointed by General Hershey to the Engineering Sciences Advisory Committee and has been attending the committee meetings at which the recommendations contained in the recently published report have been formulated. The three other members of that committee are: Chairman, Stephen L. Tyler, American Institute of Chemical Engineers; Alex C. Monteith, Westinghouse Electric Corporation; and Carl R. Soderberg, Massachusetts Institute of Technology.

Fellowships Are Available at Stanford University

Stanford University announces that financial assistance, in the form of fellowships and assistantships, is available to a substantial number of well-qualified graduate students who wish to work toward the higher degrees of M.S., Engineer, or Ph.D., either in physics or in electrical engineering, with a special interest in electronics.

Letters of inquiry will be welcomed. For application forms and further details, those who expect to major in electrical engineering should write to Assistantship Committee, E. E. Dept., Stanford University, Stanford, Calif. Those who expect to major in physics should write to Director, Microwave Laboratory, Stanford, Calif.

Applications should be submitted by March 15, 1951.

The Expansion of the IRE Professional Group System

The membership of the Institute may be justly proud of the healthy development of the IRE Professional Groups. It was the aim of these Groups to offer every member the opportunity to affiliate with other members interested in the same engineering specialities as himself. Thus the most advanced information in his preferred fields is rapidly made available to him. The Groups are, in fact, essentially technical societies within the IRE framework. They are an extremely important development, and warrant the interest and support of every member. A detailed description of their work follows. Each member should particularly study that part of this description dealing with fields of particular interest to him. It would be well if each member were to join the corresponding Professional Groups, if he has not already done so.

To join a Group, a member need only write to L. G. Cumming, Technical Secretary, The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., asking for details as to the necessary procedure.—*The Editor.*

In the space of two and one-half years, since the formation of the first IRE Professional Group, the growth of the Group system and the continuing interest in it have proven the worth of the Groups.

Since June 2, 1948, when the Institute approved the petition for the formation of the first IRE Professional Group, nine other Groups have been approved, and all have contributed to the Institute's program. The ten groups are, namely: Audio; Broadcast Transmission Systems; Antennas and Propagation; Nuclear Science; Vehicular Communications; Circuit Theory; Quality Control; Broadcast and Television Receivers; Instrumentation; and Radio Telemetry and Remote Control. Approval is now pending for the petition to form the eleventh, the IRE Professional Group on Airborne Electronics. Membership in the Groups is steadily increasing and bringing with it new members of the IRE. By now the total membership of the ten Groups is in excess of 8,500. Applications for Group membership may be obtained by applying to the Technical Secretary of the IRE.

The growth of the Professional Group System is evidenced by the fact that at the 1950 National Convention, five Groups sponsored symposia. All ten of the Groups are planning to participate at the 1951 National Convention and each has nominated a representative to serve on the Technical Program Committee for this Convention. In addition, plans are underway to set up a procedure for the systematic publication of *Transactions* of Group symposia, which will be accomplished with the co-operation of the Editorial Department. One of the Groups has recently reviewed papers for possible publication in the PROCEEDINGS. IRE Sections are kept constantly informed of the activities of the Groups by copies of "Newsletters," Conference Notes, Minutes, etc. Moreover, an account of the Group activities is published each month in the PROCEEDINGS OF THE I.R.E.

A short history of the progress made to date by each by each of the ten Groups follows:

The **IRE Professional Group on Audio** was formed with the approval of the Executive Committee on June 2, 1948. A large national meeting of the Group members took place on March 9, 1950, during the IRE National Convention. The Group has three times sponsored symposia of national scope: at the Radio Fall Meeting in Syracuse in October, 1949; at the 1950 National Convention of the IRE; and at the Radio Fall Meeting in Syracuse in November, 1950. The electroacoustics session at the 1950 National Electronics Conference in Chicago was arranged by the Group. As a service to its members, the Group began the distribution of Newsletters in March of last year. The Group has reviewed two papers for publication in the PROCEEDINGS, and on the initiative of the Group, and in co-operation with the Group, the Editorial Department at Headquarters is attempting to procure six or eight papers on Audio for forthcoming issues of the PROCEEDINGS. In December, 1949, a local Group was formed in Boston which has held technical meetings on a monthly basis since its formation. Active local Groups exist in Milwaukee and Washington, D. C., and in Detroit, the local Audio Group has been holding meetings for the past two years. Dr. Leo Beranek of the Massachusetts Institute of Technology is currently serving as the Group's Chairman. A total of 1,126 engineers are now enrolled as members of the national Group.

On July 7, 1948, the second Professional Group came into existence with the approval of a petition to form the Professional Group on Broadcast Engineers. The name of this Group was changed in August, 1949, and it is now known as the **IRE Professional Group on Broadcast Transmission Systems.** The Group is currently conducting a large membership drive aimed at the engineers in all broadcast stations and engineers concerned with the manufacture of broadcast equipment. A substantial mailing has been accomplished for the Group by IRE Headquarters which will undoubtedly result in an increase of IRE membership. The Group in the past two years was not sufficiently organized to sponsor a national symposium, but is looking toward the 1951 National Convention with a great deal of interest. A full day's program is being planned for "Broadcast Day" in March. It is anticipated that several papers presented on Broadcast Day will be submitted for possible publication in the PROCEEDINGS. In the summer of 1949 a lively program was inaugurated by the local Boston Group which has resulted in several technical meetings. Mr. Lewis Winner of Bryan Davis Publishing Company is the Group's Chairman and 808 members of the national Group are now enrolled.

On February 1, 1949, the third Group, **IRE Professional Group on Antennas and Propagation,** was formed. The national Group has been very active since the date of its formation and a great deal of time has been given by its officers toward a full and stimulating program. On October 31, November 1 and 2, 1949, in Washington, D. C., the Group, together with URSI, sponsored a highly successful symposium. On April 3 and 4, 1950, another successful symposium was sponsored with URSI, this time in San Diego, Calif. Dr. L. C. Van Atta, last year's Group Chairman, has prepared an excellent article entitled "The Role of Professional Groups in the IRE," which was published in the October, 1950, issue of the PROCEEDINGS. In accordance with the Institute's policy, abstracts of the papers presented at the Group's San Diego symposium were published in the PROCEEDINGS in August, 1950, and the Group has sponsored four papers which will be packaged as a single article for publication in a forthcoming issue. The Group's present Chairman is Mr. Newbern Smith of the National Bureau of Standards. A total of 910 members are presently enrolled in the national Group and, in addition, 300 applications for membership are pending. The Group will undoubtedly be one of the largest in the Institute.

On April 5, 1949, the Institute approved petitions for formation of three new Groups. One of these was the **IRE Professional Group on Nuclear Science.** Prior to the formation of the Group, the Technical Committee on Nuclear Studies had sponsored a nucleonics symposium jointly with AIEE. The responsibility for this annual project was assumed by the Group and on October 31, November 1 and 2, 1949, the IRE/AIEE Joint Conference on Electronic Instrumentation in Nucleonics and Medicine was held. Over 750 attended the Conference, which included an exhibition staged by the Atomic Energy Commission and a trip to the Brookhaven National Laboratory. The Group sponsored a similar Conference this year

with the AIEE on October 23, 24, and 25, again in New York City, and a technical session at the 1950 National Convention. Soon after its formation, the Group initiated a regular Newsletter which has been stimulating and informative. The former Committee on Nuclear Studies had initiated the procurement of a series of papers for the PROCEEDINGS. A dozen of these papers were published in 1948 and 1949. Upon the formation of the Group, procurement activities were assumed by the Group and four papers were published in 1950. It is expected that five more will soon be submitted. The Group's present Chairman is Mr. M. M. Hubbard of the Massachusetts Institute of Technology, and its membership totals 518.

Another Group formed on April 5, 1949, was called the Group on Vehicular and Railroad Radio Communications. The name of this Group was changed on July 12, 1950, to **IRE Professional Group on Vehicular Communications.** On November 3, 1950, in Detroit, the national Group held a full day's technical meeting on Land Mobile Communications at which eight papers were presented. A local Group has been formed in Detroit and steps are being taken toward the formation of local Groups in Chicago, New York, and Portland, Ore. Mr. Austin Bailey of the American Telephone and Telegraph Company is the present Chairman, and the membership totals 413.

The petition to form an **IRE Professional Group on Circuit Theory** was approved by the Institute's Executive Committee on April 5, 1949. The Group sponsored a technical session at the 1950 National Convention and plans to participate in the 1951 National Convention. Organization of the Group has not been completed. In addition to the 29 petition signers who are at present listed as members of the Group, Headquarters has received applications for membership totaling 994. It is expected that organization of the Group will have been accomplished by the time these notes are published. The acting Chairman for the Group is Professor J. G. Brainerd, of the University of Pennsylvania.

On July 12, 1949, the **IRE Professional Group on Quality Control** was formed by the Executive Committee's approval of its petition. The Group sponsored a technical session at the Radio Fall Meeting in 1949, and a technical session and a symposium at the 1950 National Convention. The Group again sponsored a technical session at the Radio Fall Meeting in 1950. Three papers are being published in the PROCEEDINGS on the recommendation of the Group, two of which are an outcome of the Group's symposium. Mr. R. F. Rollman of Allen B. DuMont Laboratories, Inc., is the Group's present Chairman. The membership totals 316.

The petition to form an **IRE Professional Group on Broadcast and Television Receivers** was approved by the Institute on August 9, 1949. The Group has been active in sponsoring technical meetings, including two sessions at the Radio Fall Meeting in 1949, a symposium at the 1950 National Convention, and two sessions at the Radio Fall Meeting in 1950. An article by Mr. John D. Reid, procured by the Group, was published in the October, 1949, issue of the PROCEEDINGS. The Group has an active Canadian section. There are 302 presently enrolled as members of the national Group, and 612 pending applications for membership. The Group's Chairman is Mr. Virgil Graham, of Sylvania Electric Products Inc.

January 31, 1950, saw the inauguration of the **IRE Professional Group on Instrumentation.** The Group sponsored a symposium on Improved Quality Electronic Components in co-operation with AIEE and RMA in Washington, D. C., on May 9, 10, 11, 1950. The Group sponsored the High Frequency Measurements Conference in co-operation with the Joint IRE/AIEE Committee on High Frequency Measurements which was held in Washington, D. C., on January 10, 11 and 12, 1951. The Group's Vice-Chairman, Mr. H. L. Byerlay, has prepared a short article entitled, "Organizing the Professional Group on Instrumentation at the Section Level." Copies of this article may be secured from IRE Headquarters. There exists the nucleus of a local Group in Detroit, but no action has yet been taken to set up local Groups. The Group's Chairman is Professor Ernst Weber, of the Polytechnic Institute of Brooklyn. Its membership totals 1,362, the largest of all the Groups.

The petition for the formation of the **IRE Professional Group on Radio Telemetry and Remote Control** was approved by the Institute's Executive Committee on July 12, 1950. Members have been appointed to the Administrative Committee and a meeting will soon be called to discuss plans for participation in the 1951 National Convention. The Group has not yet had an opportunity to carry on a membership drive, so at the present its members consist of the twenty-seven petition signers.

Professional Group Notes

The Committee on Professional Groups has drafted revisions to certain pages of the Manual for Professional Groups in order to outline a tentative plan for the publication of papers resulting from Group-sponsored symposia and technical meetings. The proposed plan will provide the Groups with a systematic, and effective manner of publishing the *Transactions* of their conferences and will facilitate the publication of several of these papers in the PROCEEDINGS OF THE I.R.E. . . . **A petition to form an IRE Professional Group on Airborne Electronics** has been received at Headquarters from George Rappaport of the Dayton Section. It is hoped that the Institute's Executive Committee will be able to act on the petition at its February meeting. If it is approved, the future Group will sponsor the "1951 National Conference on Airborne Electronics" in co-operation with the Dayton Section of the IRE. The Conference will be held on May 23, 24, and 25, 1951. The success of a similar conference held in 1950 lead a group of people to petition for the Professional Group. . . . **The regular joint spring meeting of the U.S.A. National Committee of the International Scientific Radio Union (URSI) and the IRE Professional Group on Antennas and Propagation** will be held April 16, 17, and 18, 1951, at the National Bureau of Standards in Washington, D. C. Information regarding the meeting may be secured from Newbern Smith, Chairman of the Group, at the National Bureau of Standards, Washington 25, D. C. . . . **The IRE Professional Group on Audio has distributed Newsletter No. 4 to** its members. The letter contains abstracts of five papers presented by the Group at the Radio Fall Meeting and at the Boston and Milwaukee Sections. A ballot for election of new officers and Administrative Committee members was included with the letter. . . . The Administrative Committee of the **IRE Professional Group on Broadcast Transmission Systems** held a meeting on December 1, 1950, at which plans for Broadcast Day at the 1951 National Convention were discussed. Two hundred applications for membership in the Group have been received as a result of the Group's recent mailing to chief engineers of broadcast stations and manufacturers of broadcast equipment. . . . **Members of the initial Administrative Committee of the IRE Professional Group on Circuit Theory have been appointed.** J. G. Brainerd, E. A. Guillemin, and W. N. Tuttle will serve on the Committee for a three-year term; W. E. Bradley and J. M. Petit for a two-year term; and R. L. Dietzold, C. H. Page, and J. R. Ragazzini for a one-year term. Professor Brainerd is Acting Chairman of the Group. . . . **The IRE Professional Group on Quality Control has** begun implementation of its plan outlined in these Notes in the January issue of the PROCEEDINGS to establish facilities to assist the IRE Sections in preparing local programs and securing outstanding authorities as speakers for meetings. Questionnaires have been sent to all Sections and these will be filled in to indicate the Section's interest or needs for speakers on quality control in the electronics field. The Group's first Newsletter has been mailed to the members. It is planned that a Newsletter will be distributed bi-monthly and will include information relevant to the general subject of quality control, abstracts of papers, and news of Group activities. . . . **The recently formed IRE Professional Group on Radio Telemetry and Remote Control has** held its first Administrative Committee meeting. In order to secure members for the new Group, a card announcing its formation has been mailed to eight of the IRE Sections which will be most interested in the Group. . . . **During the past few months Headquarters has received word of plans on the part of various IRE members which may** lead to petitions for several new Professional Groups. The fields in which interest has been expressed in recent communications are: basic science, electronic computers, engineering management, industrial electronics, information theory, and piezoelectricity.

Industrial Engineering Notes[1]

Mica Report Available

The Office of Technical Services has released a transcript of a round-table discussion by industry-military representatives on mica and mica substitutes. Subjects covered in the report include solid-state synthetic mica; World War II mica research; miniature capacitors; cellulose esters; glass and vitreous enamel insulating materials; dimensionally stable ceramics; the use of bloc talc; integrated mica; and new use of mica. Copies of the transcript PB 101 142 may be obtained for 50 cents at the OTS, U. S. Department of Commerce, Washington 25, D. C.

RTCM Releases Paper on Marine Standardization

The Radio Technical Commission for Marine Services has released a detailed paper on the "Consideration of the Difficulties Which Would Result from a Failure to Standardize Internationally Upon the Use of FM for VHF Marine Communications, Including an Analysis of the Advantages of FM over AM for Such Use."

The RTCM report deals with two specific subjects: (1) the operational difficulties which would be brought about by a dual system (FM in Region 2, AM in United Kingdom or other areas), and (2) the advantages which would result from international standardization upon the use of FM.

Copies of the report may be obtained from the Executive Secretary, R. T. Brown, Federal Communications Commission, Washington 25, D. C.

FCC Actions

The FCC has issued detailed statistics on the financial operations of AM and TV stations in 1949. The FCC tabulations, copies of which may be obtained from the Secretary of the Federal Communications Commission, Washington 25, D. C., showed that AM stations had total revenue of $413,784,633 and total income of $56,262,915. The four TV networks and their 13 owned and operated stations together with 85 other video outlets had revenues of $34,329,957 and expenses of $59,591,906 for a total broadcast loss of $25,261,950 in 1949. The FCC estimated the investment in TV broadcast stations after depreciation of $44,941,469.

Television News

A long list of prominent educators have asked the FCC to set aside vhf-uhf television frequencies and reserve them "indefinitely" for noncommercial educational purposes. Theme of the educators' contentions was that educational broadcasting could not compete with commercial TV broadcasting either for audience or for frequencies, but that the public interest required the setting aside of TV channels for educational purposes.

Credit Regulation Is Amended; Distributors Protest Ruling

The Federal Reserve Board has announced that it is amending its recently established Regulation "W" which controls consumer credit.

The revised regulation, which increases the required down payment on a TV or radio receiver from 15 per cent to 25 per cent, immediately drew the fire of radio and television distributors and dealers. The revised rules also would reduce the maximum maturity for such purchases from 18 to 15 months and lower the controlled sale from $100 to $50.

Controls

The National Production Authority has issued an order which, in effect, sets aside 10 per cent of the zinc produced for military use. The order was issued, NPA said, "to provide for the equitable distribution of of priority rated orders among all producers and fabricators of zinc . . . ." Two orders affecting the use and distribution of copper were issued by the National Production Authority in efforts to "assure copper supplies for the expanding rearmament program." The NPA order (M-12) limiting nondefense use of copper was milder than had been expected by observers. However, NPA copper officials told RTMA that the current copper situation is worse than it was at any time during World War II. This week's orders were described as the initial step. We are "merely getting our feet wet," one NPA official stated. . . . The National Production Authority has issued an order requiring tin inventories to be held to a 60 days' supply. NPA also called for reports on inventories, receipts, consumption, imports, and distribution of tin. Under the new reporting requirements, all persons having 1,000 pounds or more of pig tin in their possession or under control on the first day of the month must file a report. . . . The National Production Authority has issued an order restricting the use of aluminum for nondefense purposes to 65 per cent of the average quarterly use of the metal during the first six months of this year. The NPA directive became effective January 1. In December of this year, NPA said, users of aluminum will be permitted 100 per cent of their average monthly consumption in the first months of 1950.

Mobilization

John D. Small, former vice-president and executive assistant to the president of Emerson Radio and Phonograph Corp., was recently sworn in as Chairman of the Munitions Board. He was named to his new post by President Truman.

FCC Authorizes Tests of Subscriber-Vision System

The FCC has granted the request of the General Teleradio, Inc., licensee of WOR-TV, New York, N. Y., for special temporary authority to test, under certain expressed conditions, the "Skiatron Subscriber-Vision System" during the nonregular hours of operation of that station.

The "subscriber-vision" system of Skiatron Corp. sends a coded "scrambled" TV picture over the air. It can be received only by a receiver equipped with a special decoder. The system employs a coder unit at the camera and another at the receiver. The standard TV synchronizing pulses are fed into a small kinescope tube and projected onto a photoelectric cell, the output of which is used to supply the final synchronizing pulse.

FCC TV Network Restrictions Opposed by TV Broadcasters

About forty petitions were filed in connection with the FCC's proposal to adopt rules that would require TV stations in areas served by less than four stations to use a minimum of program time from each of the four television networks.

Majority of the petitions opposed the FCC proposal to limit one-station areas to the use of only two hours of a single network's programs in the afternoon and nighttime. Most of the petitions also recommended that full-scale hearings be held before the FCC takes any final action.

Nearly Six Million TV Tubes Are Sold in Six Months' Period

Television receiver manufacturers purchased 5,934,391 TV picture tubes in the first 10 months of 1950, according to reports to RTMA. October tube sales to set manufacturers totaled 848,387 units valued at $23,513,590, compared with 764,913 tubes valued at $20,423,353 in September.

The trend to large-type cathode-ray tubes continued in October with tubes 16 inches in size and larger representing 92 per cent of the month's sales. Rectangular tubes comprised 58 per cent of the October sales to set producers.

Total sales of all types of cathode-ray tubes, including oscillographs, camera pickup tubes, etc., in October amounted to 947,872 units valued at $26,206,183 and brought the ten months' total to 6,378,210 tubes valued at $165,611,700. . . . Television receiver production in eleven months of 1950 aggregated 6,529,615 sets, according to preliminary industry estimates. RTMA's estimates represent production by member and nonmember companies. November radio and television production each dropped eight per cent below the previous month's output. TV sets produced in November numbered 752,005 units and radio receiver production amounted to 1,304,094. Radio receivers, including home sets, auto and portables, manufactured in the 11-month period, totaled 12,785,917.

[1] The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of November 17, November 24, December 1, December 8, published by the Radio-Television Manufacturers Association, whose helpful attitude is gladly acknowledged.

IRE People

Rudolph Goldschmidt, eminent radio engineer who originated one of the first high-frequency alternators which was named for the inventor, died in November at the age of 74 in England.

Dr. Goldschmidt, who was born on March 9, 1876, at Mecklenburg, Germany, received his engineering education at the Technical University of Darmstadt. Following his graduation he served as an assistant professor.

The inventor of reflex alternators built for powers up to 100 kilowatts, Dr. Goldschmidt also invented the receiver instrument known as the tone wheel which receives continuous waves on a musical note before heterodyning.

He was president of his own company, The High Frequency Machine Company, at Berlin, during the years 1912 to 1923, and was at one time a member of the IRE.

Later Dr. Goldschmidt became an independent inventor in Berlin, and designed mechanical means for changing rotating into reciprocating motion. He also invented an electromechanical banner utilizing these means.

Thomas McLellan Davis (M'28–SM'43), head of the Radio Techniques Branch of Radio Division 11 of the Naval Research Laboratory, Washington, D. C., recently will have completed 40 years of military and civilian service in the Navy.

THOMAS McL. DAVIS

A pioneer radio engineer, Mr. Davis was born in Farmington, Maine. His naval career began with his enlistment on October 21, 1910. After a course at the Navy Electrical School, he was assigned as radio operator on the staff of Admiral Ward on the *USS Florida*, with similar duty subsequently in New York Navy Yard. From 1915 to July, 1919, he was on radio duty on Atlantic Fleet flagship, *USS Wyoming*, which was attached to the British Grand Fleet in the latter part of World War I.

He transferred from the *Wyoming* at Scapa Flow to U.S. Fleet flagship *Pennsylvania*, from which he conducted a school in British radio procedure for the radio personnel of ships stationed at New York Navy Yard. After similar duty at Norfolk, he became assistant to the Radio Officer of Washington Navy Yard's Ratio Test Shop in October, 1919. A year later he converted to civilian status at the Yard, becoming the ranking civilian in charge of the engineering and test phases of Radio Test Shop activities (except transmitters) a year prior to his transfer to NRL.

In September, 1923, Mr. Davis joined the radio staff at the Laboratory. As Head of the Radio Receiver Section, he became responsible for research, development, and design of nearly all military types of radio receiving equipment, including in addition radio direction-finder research in the period 1933 to 1940. Since 1947, he has been head of the Radio Techniques Branch, which was formed by merger of the former Receiver Section with the Radio Measurements and Components Section, encompassing a broad field of research and development in electronics.

The Navy's radio receiving equipment for the past three decades has been outstanding in reliability and performance. A major factor in this has been Mr. Davis' keen appreciation of naval radio problems and his ability to provide highly successful engineering solutions thereto.

Mr. Davis is a member of the American Physical Society and was Chairman of the Washington Section of the IRE in 1934. He served as representative to the D. C. Council of Engineering and Architectural Societies from 1935 to 1948.

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Carl L. Frederick (M'45) of Chevy Chase, Md., was recently awarded the degree of Doctor of Science by Nebraska Wesleyan University, Lincoln, Neb. The degree was recommended by J. C. Jensen, professor of physics, and was presented by Carl C. Bracy on the occasion of his installation as Chancellor.

The citation reads as follows: "Carl Leroy Frederick—Scholar and leader in scientific research, a physicist of high repute devoting time and skill to research in and for the development of means of defense from those who would destroy our democratic way of life, an alumnus of whom your alma mater is justly proud—upon recommendation of the faculty and by vote of the Board of Trustees, I am highly honored to confer upon you the Degree of Doctor of Science with all the rights, honors, and privileges of this degree."

Dr. Frederick delivered an address before a joint meeting of Sigma Pi Sigma and Phi Kappa Phi, honorary fraternities, on the subject, "Training for a Scientific Career."

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Thomas G. Banks, Jr., (A'39) has been appointed to the newly created position of Director of Research and Development at Gates Radio Co., Quincy, Ill. Prior to accepting the new position, Mr. Banks served as a Gates sales engineer for the Oklahoma-Kansas territory.

Previously he owned and managed his own broadcast station. Mr. Banks is a recognized broadcast consultant before the FCC.

Robert Charles Woodhead (A'37–M'46) died suddenly on November 8 at the age of 36. Born at Edmonton, Alberta, Canada, Wing Commander Woodhead received his engineering education at McGill University, Montreal, from which he was graduated with the degrees of B.Sc. and B.Eng.

From 1935 until World War II, he was employed in an engineering capacity by the Bell Telephone Company of Canada. Joining the Royal Canadian Air Force shortly after the outbreak of war, he was employed for a period as a navigator, and then in the Telecommunication Branch.

Retiring from the service at the cessation of hostilities to resume employment with the Bell Telephone Company, he rejoined the service in 1946 and was appointed to permanent commission in the Telecommunication Branch of the Royal Canadian Air Force. At the time of his death, he held the appointment of Director of Telecommunication Engineering, RCAF, Headquarters, Ottawa, Canada.

Wing Commander Woodhead was largely responsible for the basic engineering work connected with the domestic and overseas radio teletypewriter circuits activated by the Royal Canadian Air Force.

He also led, for a period, a joint service group which engineered an automatic tape relay which formed the backbone of the National Defense Communication System in Canada and into which RCAF major radio teletypwriter circuits were integrated.

More recently, he was responsible for the engineering decisions leading to the selection of all telecommunication equipment adopted by the RCAF, and for the general technical policy governing its use.

Marvin Hobbs (A'35–M'41–SM'43) was recently appointed Chief of the Electronics Division of the Munitions Board and Government Chairman of the Electronics Equipment Industry Advisory Committee. He has served with the Munitions Board, which is a part of the Office of the Secretary of Defense, since May, 1950, when he was appointed Deputy Director of the Secretariat of the Joint Electronics Committee.

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Norman Snyder (M'43–SM'43) of the International Standard Electric Corp., a subsidiary of the International Telephone and Telegraph Corp., is at present engaged in company activities in Saudi Arabia.

IRE People (Continued)

Hector R. Skifter (A'31–M'36–SM'43–F'51), president of Airborne Instruments Laboratory, Mineola, L. I., N. Y., has announced that a group of executives and employees of Airborne Instruments Laboratory in association with Laurance S. Rockefeller and certain of his associates, and with the American Research and Development Corporation, Boston, Mass., have purchased the entire capital stock of the Laboratory from Aeronautical Radio, Inc.

Hector R. Skifter

The transaction provides for the three groups to share nearly equal ownership of the research and development laboratory organized in 1945 as an outgrowth of three World War II laboratories associated with Columbia University, Harvard, and the Massachusetts Institute of Technology.

Management of AIL will remain in the hands of Mr. Skifter, as general manager and president; D. M. Miller, vice-president; and C. May, secretary-treasurer. The new Board of Directors is comprised of Stuart N. Scott, Joseph Powell, Jr., Georges Doriot, Randolph B. Marston, Harper Woodward, John N. Dyer, Mr. Miller, and Mr. Skifter.

❖

J. Gilman Reid, Jr., (A'50–M'50–SM'50) has been appointed Chief of the Electronics Division of the National Bureau of Standards. He has been Chief of the Engineering Electronics Section since January 3, 1939. For the present he will continue to act in this capacity, in addition to assuming duties as division chief.

In 1937 Mr. Reid joined the staff of the National Bureau of Standards as a member of the Heat and Power Division. In 1941 he became a project engineer for the uranium project at the Bureau, working on the design and development of electrical and electronic control equipment for special process control in isotope separation. From 1943 to 1944 he worked on power supply systems for radio proximity fuzes, and from 1944 to 1946 carried a major fuze project through final development and production phases. From 1946 until his appointment as Chief of the Engineering Electronics Section, he was chief engineer for the Electronic Instrumentation Laboratory, working on special instrument systems for the Navy's Bureau of Ships and Bureau of Aeronautics, the Treasury Department, and other laboratories within the NBS.

Mr. Reid attended the University of Mississippi, where he received the B.A. degree in physics in 1931 and the M.A. degree, also in physics, in 1933. From 1933 to 1936 he was a member of the staff of the Museum of Science and Industry in Chicago, and from 1936 to 1937 was on the teaching staff of the RCA Institute, instructing in physics, electronics, and mathematics.

He is a member of the American Physical Society and of the American Institute of Electrical Engineers. He has represented the Bureau on a number of society committees concerned with various phases of electronics and instrumentation.

❖

M. A. Acheson (A'36–M'37–F'41), has been transferred to the staff of E. Finley Carter, as vice-president in charge of engineering of Sylvania Electric Products Inc., in New York, N. Y.

M. A. Acheson

Mr. Acheson joined the engineering staff of Sylvania Electric in 1934. From 1935 to 1942 he supervised development engineering for radio receiving tubes, including portable types. During the war he directed Sylvania's development of proximity-fuze tubes for the Navy Bureau of Ordnance and received an award for exceptional service from the Bureau.

Appointed as manager of research and development in 1942, he served as manager of the Advanced Product Development Department of Sylvania's Central Engineering Laboratories, until he was named chief engineer for the radio tube division in 1948.

Previously, Mr. Acheson had been a member of the research staff of the General Electric Company, where he specialized in the design and development of high-power radio transmitting tubes and broadcast transmitters.

He holds approximately thirty patents on radio transmitter circuits, water-cooled power tubes, and tubes for television transmission.

❖

B. Richard Teare, Jr., (A'41–SM'45–F'51) has been appointed Dean of Graduate Studies in the Carnegie Institute of Technology College of Engineering and Science. He is also head of Carnegie's electrical engineering department and Buhl professor of electrical engineering. He will remain in these posts.

He joined the Carnegie faculty in 1939, and was named Buhl Professor in 1943. He worked toward the establishment and organization of a graduate program in the electrical engineering department, and became head of the department in 1944.

Dr. Teare has been secretary and vice-chairman of the committee on graduate degrees for the College of Engineering and Science. As Dean of Graduate Studies he will now be chairman of this committee.

As chairman of the basic course committee, and as an active teacher, Dr. Teare has contributed to the development of the Carnegie Plan of Professional Education.

In 1947, he received the George Westinghouse Award in recognition of his "distinguished contributions to engineering education."

A Fellow in the AIEE, Dr. Teare was chairman of the Institute's Committee on Education from 1948 to this year. He has been chairman of the Graduate Study Division of the American Society for Engineering Education, and is currently chairman of the Society's Division on Relations with Industry.

❖

Leister F. Graffis (S'43–A'45) has been appointed chief field engineer of the Bendix Radio Division of Bendix Aviation Corporation.

In his new capacity, Mr. Graffis will be in charge of field engineering service, maintenance training programs, and will direct the activities of technical representatives of the Bendix Radio Division who will assist government agencies in installation and maintenance. Personnel of his group will be assigned to work in many parts of this country, as well as outside the U. S. borders.

Prior to his association with the field engineering group, Mr. Graffis was with the technical publications department. Earlier he was assistant service manager of the television and broadcast receiver sales department.

Before coming to Bendix he was a Lieutenant Commander in the Navy, and was engaged in the setup of search radar equipment on the west coast and the supervision of the preparation of instructional material for all radio technician schools in that area.

He attended North Dakota State School of Science, Colorado State College, and completed radar courses at Harvard and the Massachusetts Institute of Technology. He served overseas for two years, handling the maintenance and installation of electronic equipment in military aircraft.

❖

F. C. Cahill (S'38–A'40–SM'45) supervisor of the Receiver Section at Airborne Instruments Laboratory, Mineola, L. I., N. Y., has been named supervising engineer of a combined engineering group to be known as the Radar Section. The new group, a combination of the former Receiver and Radar Sections, now comprises about 40 engineers.

Associated with Mr. Cahill in the technical operation of the new Radar Section will be assistant supervisors **Richard N. Close** (A'45), **Matthew T. Lebenbaum** (A'42–M'46–SM'47), and **William R. Rambo** (S'39–A'40–SM'46).

Mr. Cahill has been with AIL since November, 1945. He was experienced in designing commercial radio receivers and radar systems prior to joining the Radio Research Laboratory in 1942.

Books

Electrical Communications by Arthur L. Albert

Published (1950) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 581 pages+11-page index+ix pages. 428 figures. 9×6. $6.50.

This is the third edition of a work, the first edition of which appeared in 1934, and the second in 1940. The current edition has been thoroughly revised and brought up to date.

The book covers the entire field of communication, although television is treated more briefly than one might expect in view of its importance in the industry. The first portion of the book deals with the fundamentals of acoustics, electroacoustic devices, networks, lines, cables, waveguides, and various electronic devices, including recently developed ones such as the transitor. Later chapters deal with telegraph, telephone, and radio systems. The material on dial telephone and radio systems has been greatly expanded.

This book is primarily intended for use as a college text, and it should serve admirably to acquaint the students with important circuit principles, transmission systems, and terminology encountered in communication engineering. It should also be useful to one who is interested in the subject as a whole, rather than any specialized branch. The text is arranged so that the topics can be understood, even though the mathematics be ignored.

The book is thoroughly indexed, and the words and phrases listed in the index will be found printed in boldface type on the corresponding pages of the text. This is a great time saver in using the index.

Review questions have been added to the chapters, and the problems have been revised and increased in number. Extensive and modernized lists of references are given at the end of each chapter.

C. O. Mallinckrodt
Bell Telephone Laboratories, Inc.
Murray Hill, N. J.

Principles and Applications of Waveguide Transmission by George C. Southworth

Published (1950) by D. Van Nostrand Co., Inc., 250 Fourth Ave., New York, N. Y. 657 pages+xi pages+13-page index+465 figures. 6½×9¼. $9.50.

This is an up-to-date book dealing with a new art which has had a spectacular growth during the past 15 years. During all of this period of rapid growth, the author has held a position of preeminent leadership in the development of this art.

The book appears to be particularly suitable for use as a college textbook and as a reference book of great value to practicing engineers and technicians.

Two different orders of presentation are followed. The first, which is quantitative, follows the conventional theories of electricity as they apply to lumped circuits, transmission lines, waves in free space, and to waveguides. This approach includes the basic mathematical treatment of the subject. The second approach, which is largely qualitative, may be regarded as a verbal interpretation of the first, serving to develop a physical concept of the various phenomena involved.

The general background and theory are presented in the first 178 pages of text, following which, numerous applications and waveguide devices are described, including their characteristics and design factors. Included in these are transmission lines, impedance matching devices (transformers), microwave filters, balancers, amplifiers, oscillators, antennas, resonators, microwave measurements, receiving methods, radar applications, and many others.

The paper and print are of excellent quality and the presentation is enhanced by an exceptional number (465 figures) of clear-cut diagrams, curves, drawings and photographs.

H. O. Peterson
RCA Laboratories, Inc.
Riverhead, N. Y.

Frequency Modulated Radar by David G. C. Luck

Published (1950) by McGraw-Hill Book Co. Inc., 330 W. 42 St., New York 18, N. Y. 437 pages+27-page index+xviii pages. 118 figures. 6×9. $4.00.

This book is concerned primarily with the results of a Navy-sponsored research and development program at RCA in the field of frequency-modulated radar, covering a period from 1938 through the war years. The manuscript was originally prepared as a final report on a Navy contract and expectedly, therefore, considerable emphasis is placed on the theory and design of a few specific equipments. However, since the general treatment of the subject successfully transcends specific equipments, this book represents a distinct contribution to the literature, since in most radar treatises, frequency modulation systems are passed over all too briefly.

The author has done an excellent job in presenting his material in a thoroughly readable fashion using a clear, simple, and understandable style, markedly devoid of the stilted phrases and cliches often present in many technical books. A thorough knowledge of the techniques of radio engineering is assumed, together with some familiarity with pulse radar and servo mechanisms.

The book is aptly divided into two parts. Part One is devoted to the development of the theory of frequency-modulated types of systems and to discussions of various techniques used to describe the various systems developed by RCA. Particular emphasis is given to techniques that have been developed in connection with the fabrication of operable systems. The author also presents information on several developmental systems which were investigated, but not carried to completion as production items. Finally, there is included a chapter on multiple target systems in which performance comparisons are made between frequency-modulated and pulse-type radar systems. The author outlines some of the limitations as well as advantages of both types of systems.

In general, it is the conclusion of the reviewers that this book is timely, and further that it provides the electronic engineer with excellent background on the subject of frequency-modulated radar with particular emphasis on unclassified aspects of airborne fire-control and bombing systems—subjects obviously difficult to discuss in full completeness because of security requirements.

Harold A. Zahl
James T. Evers
Signal Corps Engineering Laboratories
Fort Monmouth, N. J.

Super-Regenerative Receivers by J. R. Whitehead

Published (1950) by Cambridge University Press, 51 Madison Ave., New York 10, N. Y. 157 pages +3-page index+xiii pages. 77 figures. 5½×8½. $4.75.

The author brings to his writing a wealth of experiences in direct contact with the experimental development and continuing philosophical study of one of the most unusual of all radar sets—the Mark III IFF transponder. After pioneering work in England, this super-regenerative receiver and pulse transmitter was engineered and put in large-scale production in U. S., then installed in every allied seacraft and aircraft for protection against friendly fire. Its reliable, unattended operation as an adjunct to diversified types of radar, has forever dispelled the prejudice against super-regeneration as being tricky or undependable.

After an adequate introduction to orient the reader, the author proceeds with a skillful theoretical presentation of certain special topics under his subject. Each is presented with adequate mathematical derivations and formulas, mostly relying on techniques no more advanced than the Fourier integral.

The outstanding contribution of this work is the formulation of the frequency response of a super-regenerator, which had been neglected in the few excellent studies before the war. Here the author has adeptly applied the basic principles and design formulas developed by this reviewer in Hazeltine Laboratory Reports* during the engineering of the Mark III IFF, and duly credited in other reports and publications of the author and his British associates.

Spectacular skirt selectivity is obtained by a quench wave form which provides "slope-controlled" frequency response as distinguished from "step-controlled." The properties of each are formulated and presented clearly by diagrams showing their physical significance.

The emphasis is on the "linear mode"

* H. A. Wheeler, "Superregeneration formulas," Hazeltine Report 1457W, January 5, 1943. Same, Part II, 1457 aW, February 16, 1943. (Formulas for gain and bandwidth in cases of conductance step and conductance slope.)

J. F. Craib, "Theory of superregeneration," Hazeltine Report 1506W, May, 1943. (Tests of gain and selectivity, verifying formulas in case of conductance slope.)

H. A. Wheeler, "A simple theory and design formulas for super-regenerative receivers," Wheeler Monographs, no. 3, June, 1948.

H. A. Wheeler, "Superselectivity in a super-regenerative receiver," Wheeler Monographs, no. 7, November, 1948.

and the "automatic gain stabilization" which insured the reliability of this mode in the Mark III IFF. The "logarithmic mode" is also presented clearly, with its limitations in approaching the slope-controlled selectivity.

The simplification of the presentation is achieved by dividing the quench cycle into the following periods: (1) regenerative sampling; (2) super-regenerative build-up; (3) oscillating; (4) damping. The sampling period governs the frequency response.

Some attention is paid to the noise acceptance, but the author stops short of stating some rather simple rules which express the excess noise, as compared with a simple receiver that would respond to equally short pulses.

The treatment is well illustrated by computed examples, also practical circuits and references to their war-time and post-war applications. For design purposes, the key formulas are tabulated and some are graphed. There is selective bibliography of published papers.

This work is presented on the level of an advanced student with a thorough background in the theory of communication networks. However, the diagrams and simple formulas will be instructive and helpful to the less specialized radio engineer.

HAROLD A. WHEELER
Wheeler Laboratories
Great Neck, L. I., N. Y.

Mobile Radio Handbook, First Edition, Editor, Milton B. Sleeper; Associate Editors, Jermiah Courtney and Roy Allison.

Published (1950) by FM-TV Magazine, Great Barrington, Mass. 190 pages. 249 illustrations, 26 tables. 8¾×11¼. $4.00, cloth bound. $2.00, in paper cover.

The Mobile Radio Handbook is a compilation of material of practical and technical interest to anyone considering the development of a vhf-FM radio communication system. It contains a discussion of FCC regulatory problems such as the availability of frequencies, the regulations with respect to equipment operation and operator licensing provisions. It describes in some detail many of the practical problems involved in planning, operating, and maintaining a vhf system based upon the actual experiences of some who have been active in this field. Illustrative descriptions are given of the problems involved in antenna erection, and in the maintenance of equipment as well as with the more involved problems of frequency selection, choice of equipment, siting, etc. The Handbook includes also a very good and comprehensive chapter on general FM theory.

Unfortunately much of the intensive development which has taken place in recent years in the mobile radio field has not been based upon careful over-all systems planning. Although the first chapter in the Handbook deals with basic systems planning, there could well be more emphasis on this philosophy of approach and upon the importance of providing in advance for future expansion. The Handbook suffers slightly in that it is a compilation of several separate articles, some of which apparently were not prepared specifically for inclusion in the Handbook but for presentation as magazine articles. These sections tend to be a little more specific and less objective than they perhaps should be. However, so long as the reader is interested in utilizing the Handbook as a means of acquainting himself with the general problems that may be confronted in the development, operation and maintenance of a communication system, and does not expect an exhaustive, strictly handbook-type treatment, the Mobile Radio Handbook serves a valuable purpose. It is clearly and interestingly written and very easy to read and understand.

With the extensive use now being made of vhf-FM communication systems to satisfy the operational needs of services which heretofore have not been able to use radio, and with the special problems brought about by the extensive use of such systems, there is a definite need for a handbook on mobile radio and point-to-point systems to provide guidance in the direction of proper planning and sound systems engineering. The present Handbook is a step in the right direction.

C. M. JANSKY, JR.
970 National Press Building
Washington, D. C.

Short-Wave Radio and the Ionosphere by T. W. Bennington

Published (1950) by Iliffe and Sons, Ltd., London, England. 136 pages+2-page index+57 figures. 5×8. 10s 6d.

When speaking of books, of all the branches and offshoots of the art of radio communication, the subject of radio wave propagation through the ionosphere is the most neglected. While it is true that there are few workers in the field and that the field does not lend itself towards ready experimentation, the engineer, technician, or amateur operator who feels that he would like a better and more practical knowledge of propagation finds that this is almost impossible without delving into electromagnetic theory via higher order mathematics.

Thus, it may be truly said that what this field needs is a concisely and clearly written account of the vagaries of radio wave propagation. In this reviewer's opinion, this latest volume by T. W. Bennington goes a long way toward accomplishing this very end. With greatly improved and expanded editorial and graphic content over his original volume published in 1944, this book can be readily used as a minor supplemental physics text, or simply read in one or two evenings as an interesting and well-written summation of our present state of knowledge of propagation.

The contents of the book are divided into ten chapters, each of which is then sharply subdivided by subject. Practically every phase of communication is covered in this manner and it is difficult to isolate one subject which has not been given some treatment. The entire volume is written with a complete absence of mathematics higher than fractions or simple multiplication. Each term applicable to propagation is described descriptively and with little loss of pertinent detail.

A unique part of this book was Chapter 8, which is devoted to radio amateur transmission on the high frequencies. Emphasis is placed on the inherent differences between amateur and commercial or professional practices in attempting to predict the usefulness of certain radio frequencies. The part played by sporadic-E propagation in upsetting propagation predictions is also given attention.

Considering the pains the author has gone to in literally bringing a subject down out of the clouds, it is probably very unfair to indicate pointedly the places where excess liberties in interpretations have been made. In addition, the present state is so fluid and some of our knowledge so theoretical that an author attempting to write this type of book must be granted considerable leeway. However, for example, a somewhat more up-to-date explanation of "long-scatter" would appear to be in order.

OLIVER P. FERRELL
Radio Magazines, Inc.
RASO, 121 South Broad St.
Philadelphia 7, Pa.

Radio Engineering Handbook by Keith Henney

Published (1950) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 18, N. Y. 1183 pages+12-page index+x pages. 1034 figures. 6¼×9¼. $10.00.

The fourth edition of this well-known handbook has been revised and expanded to increase its usefulness. It contains nearly twice the information of the previous edition, and has entirely new chapters on "Wave Guides and Cavity Resonators," "Electron Tubes," "Radio Aids to Navigation," "Antennas," "Receiving Systems," and "Inductance and Magnetic Materials."

The object has been to keep a working handbook with an abundance of applied information for all branches of radio engineering. Naturally, in a single volume there is considerable abridgement, but most of the chapters have excellent bibliographies attached for those who require fuller information.

This reviewer was particularly impressed by the following chapters: "Combined Circuits of L,C, and R," by W. F. Lanterman; and "V.T. Oscillators," by R. I. Sanbacher and D. C. Fielder.

Both of the above are unusually complete both in text and in references.

The chapter "Antennas," by Edward A. Laport, of 75 pages, is supplemented with a bibliography of 291 items.

"Wave Guides and Resonators," by Theodore Doreno, was excellent, well written, and terse.

"Loud Speakers and Room Acoustics," by Hugh Knowles, contains a wealth of information packed into 42 pages.

"Receiving Systems," by Charles Dean, covers the field and should constitute a valuable reference source.

"Television," by Donald Fink, is a clear exposition of basic principles which includes color basics, but not latest color systems.

"Radio Aids to Navigation," by Harry Diamond, is presented in a manner for clear, easy reading, and is comprehensive in its coverage.

In summary, the Radio Engineering Handbook is recommended as a valuable addition to any radio engineer's library.

JOHN D. REID
American Radio and TV, Inc.
Conway, Ark.

(*Continued on page 211*)

Conference on High-Frequency Measurements*

WASHINGTON, D. C.—JANUARY 10–12, 1951

SUMMARIES OF TECHNICAL PAPERS

Measurement of Frequency and Time

1. PROGRAM FOR ATOMIC FREQUENCY AND TIME STANDARDS—A SURVEY

HAROLD LYONS
(National Bureau of Standards, Washington, D. C.)

A survey will be given of the NBS program on atomic clocks and oscillators. Figures for noise-limited resolving power will be given for gases, including ammonia and oxygen, for atomic beams and nuclear quadrupole absorption. The program status will be outlined for the atomic beam clock, precision spectrograph and spectrum tables, new microwave amplifier, and multiplier tubes. A brief discussion of potential applications will include measurements of earth rotation, comparison of atomic time and the mean sidereal year, and a mm-band interferometer driven by an atomic clock for basing both time and length standards on one spectrum line.

2. IMPROVED NBS AMMONIA CLOCK

BENJAMIN F. HUSTEN
(National Bureau of Standards, Washington, D. C.)

The absorption line of ammonia is utilized to control a quartz crystal oscillator and frequency multiplier chain by means of a servo system. The servo system is designed to correct the drift of the crystal oscillator without impairing the normal short time stability of the oscillator. The control system is analyzed for stability and attainable accuracy. Minimum allowable stability figures for the uncontrolled oscillator are obtained. The theoretical limits of stability, as imposed by thermal noise and other factors, are calculated. Circuit details and performance figures are given for the complete clock and further improvements to be made indicated.

3. THE STABILIZATION OF A MICROWAVE OSCILLATOR WITH AN AMMONIA ABSORPTION LINE REFERENCE

E. W. FLETCHER AND S. P. COOKE
(Harvard University, Cambridge, Mass.)

This paper is primarily concerned with the problem of stabilizing a microwave oscillator on the 23,870-Mc absorption line of ammonia gas. First, the various causes of instability of any stabilized microwave oscillator are examined theoretically, and means for improving the stability are described and experimental results are given. Second, the feedback loop is analyzed as a servomechanism, and its design and operation are discussed. Third, the use of a spectral absorption line as a frequency reference is examined theoretically, and the design of the Cruft stabilizer is discussed.

Possibilities for further improvement of absorption-line frequency references are examined. This includes a discussion of the dependence of the feedback signal on the pressure and temperature of the ammonia and the dimensions of the waveguide cell.

4. PERFORMANCE OF OSCILLATORS FREQUENCY-CONTROLLED BY GAS ABSORPTION LINES

L. E. NORTON
(RCA Laboratories, Princeton, N. J.)

Frequency stabilization of an oscillator implies comparison of two frequencies, output and standard, with provision for output frequency correction derived from any difference between the two. At microwave frequencies molecular absorption lines are particularly useful frequency standards. Over-all stability of a frequency-controlled oscillator is specified by three things: namely, original frequency stability of the uncontrolled oscillator, stability of a standard frequency derived from an absorption line, and a multitude of factors related to comparison circuit-servo control loop elements. Stabilized oscillator performance is predicted by calculating the error magnitude or uncertainty introduced by each of these various causes. Details of these effects are described for specific circuits.

5. MILLIMETER-WAVE MEASUREMENTS

WALTER GORDY
(Duke University, Durham, N. C.)

The discussion will include methods of generating, detecting, and measuring millimeter-wave frequencies in the range of 30,000 to 150,000 Mc. A table of accurately measured spectrum lines will be given. These provide suitable frequency standards for calibration of cavity wave meters in the millimeter range. The millimeter-wave absorption of oxygen, carbon monoxide, and other gases will be discussed.

6. QUARTZ-CRYSTAL FREQUENCY STANDARDS

W. D. GEORGE
(National Bureau of Standards, Washington, D. C.)

The past and present performance of low-frequency quartz-crystal resonators and oscillators, as used in maintaining the national standard of frequency and time interval, will be summarized. Improved accuracy of the frequency standard is resulting from recent improvements in uniformity of standard time, greater reliability in temperature control and better GT crystal units and other components. Problems in connection with distribution of the standard via radio broadcasts will be discussed, e.g., accuracy as received, coverage, continuity, corrections, type of service, and simultaneous use of two or more stations.

7. HIGH-FREQUENCY CRYSTAL UNITS FOR PRIMARY FREQUENCY STANDARDS

A. W. WARNER
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

A new approach to the design of crystal units for primary frequency standard use has resulted in crystal units, in the 3-to-20-Mc frequency range, characterized by high Q and low capacitance in the series arm of the equivalent electrical circuit.

By utilizing the overtone frequency of specially designed AT-cut quartz plates, both Q and the rate of impedance change with frequency are enhanced together, and in addition the stability with time of the crystal unit is increased, due to a larger frequency determining dimension. Additional characteristics of the crystal units include small size, stability under conditions of vibration and shock, and low temperature coefficient.

Stabilities of one part in 10^8 per month have been achieved without recourse to stabilized circuits.

Measurement of Impedance

8. INFLECTION-POINT METHOD OF MEASURING Q AT VERY-HIGH FREQUENCIES

NELSON E. BEVERLY
(Sperry Gyroscope Company, Great Neck, L. I., N. Y.)

A new method of measuring Q has been developed for measurements at very-high and microwave frequencies. This method determines the inflection points of a resonance curve by a null indication, and Q is obtained only as a function of the frequency at which these null indications occur. The method is broad in application in that it can be used to locate the inflection points of lumped-constant circuits, transmission lines, and microwave cavities.

Experimental measurements have been made at vhf and microwave frequencies. The results indicate that the accuracy may be equal or better than results obtained by other known methods.

* Sponsored by The Institute of Radio Engineers, the American Institute of Electrical Engineers, and the National Bureau of Standards; and organized by the Joint AIEE-IRE Committee on High-Frequency Measurements, the IRE Professional Group on Instrumentation, and the AIEE Committee on Instruments and Measurements.

9. A PRECISE SWEEP-FREQUENCY METHOD OF VECTOR IMPEDANCE MEASUREMENT

D. A. ALSBERG
(Bell Telephone Laboratories, Inc. Murray Hill, N. J.)

The impedance of a two-terminal network is defined completely by the insertion loss and phase shift it produces when inserted between known sending and receiving impedances.

Recent advances in precise wide-band phase and transmission measuring circuits have permitted practical use of this principle. These circuits are free from zero corrections as the measuring frequency is changed, which in one specific circuit can be swept continuously from 0.05 to 20 Mc while data can be recorded automatically with accuracies up to ±0.25 per cent. Reactive and resistive impedance components are read directly from a simple graphical chart in which frequency is not a parameter. The basic principle described promises attractive possibilities in many cases of impedance measurements at still higher frequencies, where present methods are inadequate.

10. PRECISION COAXIAL RESONANCE LINE FOR IMPEDANCE MEASUREMENTS

HOWARD E. SORROWS, ROBERT E. HAMILTON, AND WILLIAM E. RYAN
(National Bureau of Standards, Washington, D. C.)
MING S. WONG
(Aircraft Radiation Laboratory, Wright Field, Dayton, Ohio)

As a part of a program to develop standards of measurements of electrical quantities at ultra-high frequencies, an instrument has been constructed with which precise measurements can be made by a resonance-curve method. Sources of error in measurements made with this "Chipman" type line are discussed, and methods are presented for evaluating the magnitudes of those errors which cannot be made negligible by use of proper techniques. A theoretical analysis of the errors is presented in a form which can be applied directly to the determination of the accuracy of impedance measurements made with the line. The results of this study, and those of a similar and recent study of the slotted line, provide a direct comparison of the accuracy which can be obtained by the two methods of measurement.

11. A 2,600- to 4,000-Mc VSWR-MEASURING SET

S. F. KAISEL
(RCA Laboratories, Princeton, N. J.)
J. W. KEARNEY
(Airborne Instruments Laboratory, Mineola, L. I., N. Y.)

A VSWR-measuring unit is described which will present on an oscilloscope a quantitative picture of the VSWR looking into a waveguide or coaxial-line element over the frequency range 2,600 to 4,000 Mc. Standing-wave ratios of 1.5 to 1 or greater can easily be measured. The unit consists of a mechanically-swept reflex klystron oscillator, which tunes from 2,200 to 4,600 Mc continuously, as an rf signal source, feeding a waveguide Magic Tee as the measuring element. The unbalance of the Magic-Tee bridge is presented on an oscilloscope. Means are provided for calibrating the oscilloscope directly in terms of voltage standing-wave ratio.

12. MEASUREMENT OF WAVEGUIDE AND COAXIAL LINE IMPEDANCES WITH A CIRCULAR WAVEGUIDE

ARTHUR E. LAEMMEL
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

A pickup loop is rotated in a circular waveguide which is joined to the waveguide or coaxial line containing the unknown impedance. The rotary variation of voltage induced in the loop can be used to measure impedance in the same way that the linear variation of voltage picked up by the probe of a slotted section is used, since both patterns are the same under suitable conditions. Some of the advantages of the rotary device are smaller size, constancy of wavelength (always being one rotation), and ease of mechanical drive.

13. SURVEY OF MICROWAVE DIELECTRIC TECHNIQUES FOR SMALL LIQUID AND SOLID SAMPLES

GEORGE BIRNBAUM
(National Bureau of Standards, Washington, D. C.)

This paper surveys the theory and application of cavity systems suitable for measuring the microwave dielectric properties of liquids and solids in the form of small-diameter cylindrical samples. The particularly simple equations, connecting dielectric constant with resonance frequency and loss factor with Q, for such samples, suitably located in a TM_{010} cylindrical cavity or a TE_{10} rectangular cavity, are discussed. Simple, yet sensitive apparatus for measuring these cavity parameters are described. Data obtained by several observers in determinations of the dielectric properties of threads and ionized gases, as well as liquids and solids, are discussed to demonstrate the accuracy and limitations of these methods.

Demonstration Lectures

14. MICROWAVE SPECTROSCOPY WITH APPLICATION TO CHEMISTRY, NUCLEAR PHYSICS, AND FREQUENCY STANDARDS

L. J. RUEGER, R. G. NUCKOLLS, AND HAROLD LYONS
(National Bureau of Standards, Washington, D. C.)

A Stark-modulation type of microwave spectrograph will be demonstrated by displaying the absorption lines of ammonia gas on a projection oscilloscope. Nuclear quadrupole hyperfine structure lines will be shown, the Stark effect, pressure broadening of the lines, and possibly power saturation. It will be seen how applications to chemical analysis, chemical reaction rates, isotope analysis, and nuclear physics are possible. The usefulness of the lines as invariant frequency standards will be seen, and the possible application of precision designs of spectrograph to measurements of the rotation of the earth using these lines will be pointed out.

15. RECORDING ATMOSPHERIC INDEX OF REFRACTION AT MICROWAVES

GEORGE BIRNBAUM, S. J. KRYDER, AND R. R. LARSON
(National Bureau of Standards, Washington, D. C.)

The recording microwave refractometer is an instrument which measures and records minute differences in frequency between a test cavity and a reference cavity. The operating principle and the applications of this instrument for research and industrial control work will be briefly described.

The use of this instrument to measure and record fluctuations in the refractive index of the atmosphere will be demonstrated. The output meter of the instrument will be projected on a screen, and the variations in meter readings will be noted as someone breathes into the test cavity or a moist blotter is held near it. Recordings of fluctuations in the refractive index of the atmosphere obtained at the National Bureau of Standards will be projected, and the significance of these records will be briefly discussed.

16. MEASUREMENT OF MICROWAVE FIELD PATTERNS USING PHOTOGRAPHIC TECHNIQUES

W. E. KOCK
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

A description of a mechanical scanning method for photographically displaying the space patterns of microwaves and centimeter wavelength sound waves will be followed by a demonstration of refraction, diffraction, and focusing of these waves by iterative metallic structures. Photographs of a large variety of field patterns will be shown. The position of the individual wave crests and the direction of wave motion can be indicated by the addition of a constant amplitude signal. Simultaneous focusing of sound waves and microwaves by the same lens will be demonstrated.

Measurements of Power and Attenuation

17. ABSOLUTE MICROWAVE POWER MEASUREMENTS

A. C. MACPHERSON AND D. M. KERNS
(National Bureau of Standards, Washington, D. C.)

Work done at the NBS concerning bolometers and calorimeters as devices for

precise, absolute measurement of microwave power will be reviewed. Special techniques developed for the determination of bolometer mount efficiency by means of impedance measurements will be described briefly. A differential microcalorimeter, developed and used for power measurement at the milliwatt level, will also be described. A summary of measurements consisting largely of cross-checks between the bolometric and the calorimetric methods will be presented, and evidence for the failure of the impedance method of determining efficiency for thermistors and the success of the method for platinum-wire bolometers will be discussed.

18. BROAD-BAND BOLOMETER DEVELOPMENT

W. E. WALLER
(Polytechnic Research and Development Co., Inc., Brooklyn, N. Y.)

Broad-band bolometers for high-frequency power measurements have been developed to cover the ranges 20 to 1,000 Mc, 1,000 to 4,000 Mc, and 4,000 to 10,000 Mc. All units have a VSWR under 1.3 over their specified operating ranges. Low-power and high-power elements, capable of dissipating 1 milliwatt and 100 milliwatts, respectively, have been made to the above specifications. The response and power handling capabilities of these elements to short pulses will be discussed.

19. CALIBRATING AMMETERS ABOVE 100 MC

H. R. MEAHL AND CHARLES C. ALLEN
(General Electric Company, Schenectady, N. Y.)

A survey is made of the progress to date in measuring current above 100 megacycles. The types of vacuum thermocouples available for ultra-high-frequency current measurement are discussed, and the several methods of calibration are reviewed. A calorimeter method and a thermistor bridge method are presented. The advantages and limitations of the calibrating methods are brought out. The electrodynamic method is particularly suited to large currents, the calorimeter method to medium currents, and the thermistor bridge method to small currents. The importance of obtaining agreement between methods that do not depend on the same principles is emphasized.

20. A MICROWAVE OSCILLOGRAPH

W. B. SELL AND J. V. LEBACQZ
(The Johns Hopkins University, Baltimore, Md.)

A cold-cathode, high voltage, single transient Rogowski oscillograph was given to The Johns Hopkins University by the Aberdeen Proving Grounds. The oscillographic chamber was redesigned to permit the direct observation of frequencies in the 10,000-Mc range. This has been accomplished in two ways: first, by velocity-modulating the beam by the E field of a waveguide, then deflecting it through a constant magnetic field; second, by using a short Lecher wire system for the transmission of the microwave energy. In this latter case, deflections due to both the E and H fields have been observed. A theory is offered which checks the E-field deflection much more closely than the predictions of Hollmann.[1] A qualitative explanation of the H-field deflection is also given.

21. PRECISION MILLIDECIBEL WAVEGUIDE ATTENUATION MEASUREMENTS

J. H. VOGELMAN
(Watson Laboratories, Red Bank, N. J.)

The precision measurement of the attenuation of four-terminal low-loss microwave structures is based on the relationship between the attenuation and the resultant standing wave at the input terminal when the structure is terminated in a short circuit. The techniques and necessary precision measurement equipment have been developed to permit accurate measurement of attenuation values between 0.01 and 0.5 db, and with decreasing accuracy down to 0.001 db. Since the resultant VSWR values are very large, the ultimate accuracy depends on the determination of the relative magnitude of the minimum with respect to the readily measurable maximum. Techniques and equipment will be described which minimize the errors due to nonlinearity of detectors, power reflections from the test sample, noise in the detector amplifiers, residual frequency modulation of signal source (long-line effects), and attenuation in measuring line. The attenuation measurement has been reduced to the measurement of two physical lengths.

22. DISSIPATIVE AND PISTON ATTENUATOR CORRECTIONS

CHARLES M. ALLRED
(National Bureau of Standards, Washington, D. C.)

Termination impedances and corrections on rated values of dissipative-type attenuators terminated in any complex impedances will be discussed. Derivation construction and application of several circle diagrams and nomographs will be presented, and experimental verification data shown for a wide range of frequencies and impedances. In addition, derivation and construction of nomographs will be presented, giving both cutoff frequency and skin penetration corrections for TE_{11} piston attenuators. Slides will be shown of the above nomographs, circle diagrams, theoretical highlights, and some of the latest NBS attenuation standard equipment.

Measurement of Transmission and Reception

23. A FIELD-STRENGTH METER FOR 600 MC

J. A. SAXTON
(National Physical Laboratory, Teddington, England)

The equipment has been designed primarily for use in the study of radio wave propagation at 600 Mc, and has a fairly wide-band intermediate-frequency amplifier (0.5 Mc centered on 30 Mc) to ensure that a received signal remains in tune during long periods of recording. As a calibrated instrument, however, it may be used for the accurate comparison of radio field strengths, or powers, over a frequency range 500 to 700 Mc. It can be operated with a continuous-wave signal, when the sensitivity (for signal equal to noise) with a dipole receiving aerial is about 35 μv/m, or with a modulated signal; in the latter case a narrow-band audio-frequency amplifier is added, and the over-all sensitivity of the equipment is about 18 db greater than with an unmodulated signal.

[1] Hans E. Hollmann, "The dynamic sensitivity and calibration of cathode-ray oscilloscopes at very-high frequencies," PROC. I.R.E., vol. 38, p. 32; January, 1950.

24. MEASURING TECHNIQUES FOR BROAD-BAND LONG-DISTANCE RADIO RELAYS

W. J. ALBERSHEIM
(Bell Telephone Laboratories, Inc., Deal, N. J.)

Adjustment and maintenance of radio relays require sensitive, yet rapid, measurements.

By rapid scanning, transmission characteristics can be traced on paper strips or cathode-ray screens. Alternating switches permit superposition of reference traces.

Frequency functions thus measurable include gain, phase, impedance, reflection coefficient, and their rates of change; time functions include amplitude and frequency modulation; amplitude functions—FM distortion caused by discrete frequency or noise modulation, by interference and by transmission characteristics.

Time and level distributions of atmospheric disturbances are recorded, and the effects of selective fading and echoes evaluated by simulating them under controlled laboratory conditions.

25. WIDE-BAND SWEPT-FREQUENCY MEASUREMENTS APPLICABLE TO TRAVELING-WAVE TUBES

FREDERICK E. RADCLIFFE
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Methods of measuring transmission and impedance of traveling-wave tubes operating in the 4,000-Mc common-carrier band are described in which the frequency is swept over a 500-Mc band. The characteristics are displayed on a standard oscilloscope.

A gain-comparator type of transmission measuring set is described in which moderate oscillator amplitude variations with frequency do not affect the accuracy of measurement. By this method transmission characteristics are measured to accuracies of about 0.25 db and return loss can be measured up to values of 40 db.

A new technique, which is useful in broad-band amplifier measurements in general, is described in which both the transmission and output impedance-versus-frequency characteristics of an amplifier delivering its normal power output to its load, are displayed simultaneously on the oscilloscope. Thus adjustments of the output

impedance of a multistage amplifier can be made while compensating the transmission characteristic in another part of the amplifier circuit.

26. MICROWAVE TECHNIQUES IN THE 28,000- to 300,000-Mc REGION

Leonard Swern
(Sperry Gyroscope Company, Great Neck, L. I., N. Y.)

This paper will survey the techniques of microwave measurements in the millimeter region. The problems involved in using short wavelengths will be discussed, and certain solutions to these problems indicated. Several examples of new millimeter-region components will be described and illustrated by photographs. These designs will be evaluated. In addition, certain new design proposals will be described. The paper's emphasis will be on techniques fundamentally similar to those effective at lower frequencies. However, new approaches to the microwave measurements problem will be discussed, among them, optical and semi-optical approaches, and the applicability of molecular resonance absorption.

27. MEASUREMENT OF CHARACTERISTICS OF CRYSTAL UNITS

L. F. Koerner
(Bell Telephone Laboratories, Inc., Murray Hill, N. J.)

Recent extensions of the useful frequency range of crystal control into the microwave region by harmonic generation, and the extension of the frequency range of crystal units in filter networks, have called for more accurate measurements of the electrical elements of the equivalent network of the crystal unit. The agreement of frequency measurements of a crystal unit in various test circuits is a function of its Q, and the ratio r of its shunt capacitance to the capacitance in the series branch of the equivalent network. Measurements of the resonant frequency of the series branch in various circuits may vary by $100\times r/2Q^2$ per cent, and the spread between the frequency of zero phase angle and the frequency of minimum impedance of the equivalent network may be double this value. If the quality of the crystal unit is maintained sufficiently high, these variations are of the order of a few cycles per megacycle and may be neglected except for precision applications, as in frequency standards. For crystal units operating at frequencies in excess of about 10 Mc, the shunt capacitance is equalized by an equivalent positive reactance to reduce these variations.

28. REFLECTING SURFACE TO SIMULATE AN INFINITE CONDUCTING PLANE AT MICROWAVE FREQUENCIES

S. J. Raff
(U. S. Naval Ordnance Laboratory, Silver Spring, Md.)

The shape of a flat conducting surface has considerable effect on the microwave reflections from it, when the source to reflector distance is of the order of magnitude of the dimensions of the reflector. An optimum shape is designed for simulating the reflection back to the source of an infinite-plane reflector. If the reflection is summed by Fresnel zones, the large effect of the shape of the surface on the reflections can be attributed to the slow rate of decrease of reflections from successive zones. Using the calculus of variations, an optimum shape is designed for a surface inscribed in a circle of 23 wavelengths diameter, 25 wavelengths from a dipole antenna. The maximum error in simulating an infinite-plane reflection was calculated to be 3 per cent. The reflection from this surface has been measured, and is within 4 per cent of the infinite-plane reflection over the range of 15 to 35 wavelengths from the antenna

Books (*Continued*)

Television Installation Techniques by Samuel L. Marshall

Published (1950) by John F. Rider Publishers, Inc., 480 Canal Street, New York 13, N. Y. 336 pages+4-page index+270 figures. 5⅜×8⅜. $3.60.

This outstanding book was written as a reference and handbook, primarily for the television service man, but it is of great interest to the experimenter, the engineer, and the service manager.

The first four chapters of the book present a well-balanced discussion of the nature of television, radio propagation antenna and transmission lines, and transmission and special antenna systems, supported by a clear-cut discussion of the essential theory, and supplemented by specific design methods and practical design information to meet almost any installation requirement.

The next two chapters entitled, "Materials and Methods used in Installations" and "High Mast and Tower Installations," cover the installation of antennas in primary service areas, with particular regard to safety and best installation practices. In the case of the high mast and tower installations, the principles of construction design formulas and data to take care of wind and ice loadings, and special design considerations are discussed in detail for most types of masts and towers.

Two chapters are devoted to problems arising from television installations and receiver adjustments in the home. Problems in connection with reflections, multiple installations, fringe area operations, television interference, TV filters, and the adjustment and servicing of sets in the home are covered in a most factual and satisfactory manner.

The last chapter, "Municipal Regulations," covers safety precautions in general, and the National Board of Fire Underwriters Bulletin 275, the municipal codes and television ordinances of many of the cities of the United States.

The Appendix includes a great deal of useful information in its ten tables and three charts pertaining to vhf TV stations on the air: Data on coaxial cables and transmission lines, data relative to the safe loads of anchors, bolts and guy cables, tape sizes, and a list of the various sizes and characteristics of cathode-ray picture tubes.

This book includes authoritative information describing the best field practices in the installation of antennas to meet practical field problems and the installation and servicing of receivers to give maximum satisfaction to the customer, and points out the importance of attention to details in installations, both in the interest of the customer and of safety.

Lewis M. Clement
Crosley Division
Avco Manufacturing Corporation
Cincinnati, Ohio

Nuclear Data is a new publication of 310 pages available from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C. The price of $4.25 a copy includes the cost of three supplements (about 60 pages each) which the purchaser will automatically receive at six-month intervals. Remittances from foreign countries must be made in United States exchange and must include an additional sum of one-third the publication price to cover mailing costs.

Nuclear Data is a valuable publication for nuclear physicists and engineers, radiochemists, biophysicists, and other workers in the rapidly expanding field of nuclear physics. It presents a collection of experimental values of half-lives, radiation energies, relative isotopic abundances, nuclear moments, and cross sections. Decay schemes and level diagrams, over 125 of which are included in the tables now ready, are to be provided wherever possible.

The National Bureau of Standards, with the assistance of the Oak Ridge National Laboratory, the Brookhaven National Laboratory, the Massachusetts Institute of Technology, and the University of California Radiation Laboratory, is making the first effort to present a continuing compilation in this rapidly developing field. The present tables and the supplements to follow are therefore designed for easy assimilation of new material in loose-leaf form.

Abstracts and References

Prepared by the National Physical Laboratory, Teddington, England, Published by Arrangement with the Department of Scientific and Industrial Research, England, and *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, and not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S. E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

ACOUSTICS AND AUDIO FREQUENCIES

016:534 **1**

References to Contemporary Papers on Acoustics—A. Taber Jones. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 643–652; September, 1950.) Continuation of 2682 of 1950.

534.213.4 **2**

On the Propagation of Sound Waves in a Cylindrical Conduit—F. B. Daniels. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 563–564; September, 1950.) "The characteristic impedance and propagation constant of a cylindrical conduit are calculated on the basis of an equivalent electrical T section. Numerical values of the results are plotted for air at 20°C, for a range of values of the independent variable which includes the region of transition from isothermal to adiabatic conditions."

534.213.4 **3**

On the Propagation of Sound in Narrow Conduits—O. K. Mawardi. (*Jour. Acous. Soc. Amer.*, vol. 22, p. 640; September, 1950.) Correction to paper abstracted in 2 of 1950.

534.231 **4**

Reaction of Medium and Acoustic Radiation Damping for a Disk—H. Braumann. (*Z. Naturf.*, vol. 3a, pp. 340–350; 1948.) In order to calculate the sound field of a disk executing small harmonic oscillations in the direction of its axis, the acoustic wave equation is formulated in bipolar co-ordinates. For the case where the medium may be regarded as incompressible the wave equation leads to the potential equation; an exact solution is given for this. The effect of the medium on the disk is equivalent to an increase of $8R^3\rho_0/3$ in its mass, where R is the radius of the disk and ρ_0 the density of the medium. For a compressible medium there is an additional effect due to radiation damping. The time average of the radiated power is approximately $8R^6\rho_0\omega^4V^2/27\pi c^3$, where ω is the angular frequency, V the maximum velocity of the disk, and c the velocity of sound.

534.231:534.133 **5**

The Sound Field of a Straubel X-Cut Crystal—E. W. Samuel and R. S. Shankland. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 589–592; September, 1950.) The streaming patterns in CCl_4 were determined, using a 7.5-Mc crystal and measuring the velocities of sugar particles whose density is nearly that of the liquid. The results indicate that the intensity of the sound radiated by a crystal of the Straubel type is very uniform across the beam.

534.231:621.395.623.75 **6**

Response Peaks in Finite Horns—C. T. Molloy. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 551–557; September, 1950.) The axial sound field of horn-type loudspeakers is calculated theoretically. The frequencies at which peaks occur in the response curves, and the dimensions of hyperbolic and exponential horns having peaks at specified points in their response curves, can be derived. A comparison between a measured and a computed axial response curve shows good agreement.

534.231.3:621.395.623.75 **7**

Resonance Characteristics of a Finite Catenoidal Horn—G. J. Thiessen. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 558–562; September, 1950.) Expressions for the impedance components of a finite catenoidal horn are derived and a comparison with similar exponential and conical horns is made. The impedance of a section of a catenoidal horn is also calculated; for the finite as well as the infinite horn, this impedance approaches that of the exponential horn as more length is trimmed from the throat end.

534.232:517.564.3 **8**

On the Extension of Some Lommel Integrals to Struve Functions with an Application to Acoustic Radiation—C. W. Horton. (*Jour. Math. Phys.*, vol. 29, pp. 31–37; April, 1950.)

534.24 **9**

On the Non-specular Reflection of Plane Waves of Sound—V. Twersky. (*Jour. Acoust. Soc. Amer.*, vol. 22, pp. 539–546; September, 1950.) Mathematical analysis of the scattering of sound by various rigid nonabsorbent nonporous surfaces consisting of semicylindrical or hemispherical bosses on an infinite plane. Exact solutions are given for the case of a single boss and a plane wave at an arbitrary angle of incidence. The theory is extended to the case of finite arrangements of bosses, neglecting secondary excitation effects. The solutions contain the usual Fraunhofer terms for a grating or lattice. The cases of finite and infinite random distributions are considered, and various extensions of the treatment are mentioned.

534.321.9:061.3 **10**

Rome Ultrasonics Convention—G. Bradfield. (*Electronic Eng.* (London), vol. 22, pp. 391–394; September, 1950.) Brief account of lectures and exhibited equipment, including commercial apparatus for many purposes. The advantages of $BaTiO_3$ ceramics for piezoelectric applications were pointed out. A method of investigating the properties of auditoriums, using models and ultrasonic frequencies, was described.

534.321.9:534.614–14 **11**

A New Method for Measuring Velocities of Ultrasonic Waves in Liquids—B. R. Rao. (*Nature* (London), vol. 166, p. 742; October 28, 1950.) A rapid, accurate method requiring only a very small quantity of the liquid and applicable to both opaque and transparent liquids over a wide range of frequencies is described.

534.374 **12**

Mathematical Analysis of an Acoustic Filter—N. Olson. (*Canad. Jour. Res.*, vol. 28, Sec. A, pp. 377–388; July, 1950.) The attenuation, phase, and impedance functions are calculated for acoustic filters constructed from conduits with a series of equal wider sections at regular intervals. Theoretical impedance and attenuation curves are shown and are confirmed by measurements on filters of circular and square section. Such filters are easy to construct and terminate and they may be used in parallel to form units of large cross section.

534.612+534.641 **13**

A Method for Measuring Source Impedance and Tube Attenuation—J. E. White. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 565–567; September, 1950.) A description of a sensitive and accurate method of determining attenuation and velocity of sound from sound-pressure measurements in gases at medium audio frequencies. By using a microphone movable along an open-ended tube, the acoustic impedance of any sound source coupled to the air column can be found if the acoustic impedance of the microphone is known. All necessary formulas are derived.

534.841/.844 **14**

Reverberation Time and Sound Power Required for Ordinary Rooms—E. de Gruyter. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 757–761; September 17, 1949. In German.) A new interpretation of Sabine's formula is given which takes account of measurements of reverberation duration made in halls with good acoustical properties. Reverberation constants are derived which characterize the acoustic quality of a room for any particular purpose and which are independent of room dimensions. A formula for the necessary acoustic power is then established by analogy of the reverberation constants with the power constants of an electrical circuit. Psychological considerations

affecting the question of good acoustics are discussed, in particular in connection with modern electroacoustic transmissions.

534.843 15
On the Acoustics of Coupled Rooms—C. M. Harris and H. Feshbach. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 572–578; September, 1950.) The problem of coupled rooms is treated as a wave-theory boundary-value problem. Experimental data for isolated modes of vibration of a coupled system confirm the theoretical formulas.

534.845 16
Parameters for Sound Transmission in Granular Absorbing Materials—M. Ferrero and G. Sacerdote. (*Nuovo Cim.*, vol. 5, pp. 551–566; December 1, 1948.) The parameters are deduced from acoustic-impedance measurements on lead shot and on sand of various grain sizes and are discussed in relation to theory previously developed (*ibid.*, vol. 4, p. 262, 1947).

534.86 17
The "Expressor" System for Transmission of Music—R. Vermeulen and W. K. Westmijze. (*Philips Tech. Rev.*, vol. 11, pp. 281–290; April, 1950.) It is claimed that in the transmission of music, either by radio or via a sound-recording system, adjustment of the degree of volume compression by a capable hand has great advantages over automatic control. But with manual control the relations between the intensity of the original music and that of the input signal of the expandor is ambiguous, and special methods have to be used to ensure that the expandor always exactly compensates the compression. In the 'Expressor' system, a pilot signal consisting of impulses is transmitted via a separate channel and causes the potentiometer of the expandor to follow continuously the movements of the compressor.

548.1:537 18
Piezoelectric Equations of State and Their Application to Thickness-Vibration Transducers—Cady. (See 99.)

621.395.61.089.6 19
American Standard Method for the Pressure Calibration of Laboratory Standard Microphones: Z24.4—1949 (Abridged)—L. L. Beranek, R. K. Cook, F. F. Romanow, F. M. Wiener, and B. B. Bauer. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 611–613; September, 1950.) The complete standard may be obtained from the American Standards Association, 70 East 45th Street, New York 17, N. Y., for 75 cents.

621.395.613.37 20
A Second-Order-Gradient Noise-Canceling Microphone Using a Single Diaphragm—W. A. Beaverson and A. M. Wiggins. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 592–601; September, 1950.) The four openings admitting sound to the two surfaces of the microphone diaphragm are so spaced and oriented that the response depends on the second differential of the sound pressure with respect to the direction of propagation. Theoretical analysis, confirmed by experimental results, shows that a better signal-to-noise ratio is obtained than with first-order-gradient microphones.

621.395.623.54.089.6 21
American Standard Method for the Coupler Calibration of Earphones: Z24.9—1949 (Abridged)—L. L. Beranek, F. F. Romanow, K. C. Morrical, L. J. Anderson, B. B. Bauer, R. K. Cook, and W. Wathen-Dunn. (*Jour. Acoust. Soc. Amer.*, vol. 22, No. 5, pp. 602–608; September, 1950.) The complete standard may be obtained from the American Standards Association, 70 East 45th Street, New York 17, for 75 cents.

621.395.625.2:621.396.933 22
An Automatic Monitoring Recorder—See 207.)

621.395.92 23
Hearing-Aid Design—A. Poliakoff. (*Wireless World*, vol. 56, pp. 274–276; August, 1950.) Among the essential qualities of a good instrument, the most important is the provision of optimum volume to suit each patient in all reasonable conditions of use. In most cases this cannot be achieved without automatic volume controls. There should be no pronounced peaks in the response curve and case noise should be reduced by mounting the microphone in rubber and making the surface of the case very smooth.

621.396.822:621.316.8 24
A Thermal-Noise Generator for Low-Frequency Tests—H. Meister. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 320–324; August 1, 1950. In French and German.) A description is given, with a circuit diagram showing all component values, of a generator providing a continuous noise spectrum from 30 cps to 15 kc. Reasons for the choice of a resistor as the primary noise source are given. The problems connected with the high amplification necessary with such a source are discussed and means for reducing instability are indicated.

621.396.822:621.385.38 25
A High-Level Noise Source for the Audio-Frequency Band—H. D. Harwood and D. E. L. Shorter. (*Jour. Sci. Instr.*, vol. 27, pp. 250–251; September, 1950.) An argon-filled thyratron is used to generate a continuous noise spectrum for audio-frequency testing. Large unwanted ultrasonic components are removed by means of a low-pass filter cutting off at 20 kc, leaving a noise spectrum with power distributed uniformly over the audio-frequency band. Where required, the power per octave band can be made uniform by adding a weighting network. The output into 600Ω is −20 db with reference to 1 mw. The generator is mains operated.

ANTENNAS AND TRANSMISSION LINES

621.315.21.017.71.029.4/.6 26
The Power Rating of Radio-Frequency Cables—R. C. Mildner. (*Trans. AIEE*, vol. 68, Part I, pp. 289–298; discussion, p. 298; 1949.) The limitations to the power that can be transmitted by various types of radio-frequency cable are considered, suitable rating factors to deal with different maximum permissible cable temperatures and different ambient temperatures are proposed, and the influence of modulation, standing waves, and load factor on the thermal and voltage rating is discussed. The principles outlined are applied to determine the ratings of three types of cable.

621.315.212:621.397.24.018.78† 27
Characteristics of Coaxial Pairs of Frequencies Involved in High-Definition Television Transmission—G. Fuchs. (*Câbles & Trans.* (Paris), vol. 4, pp. 248–254; July, 1950.) The distortion, caused by irregularities of cable impedance, which would occur in the transmission of a television signal of 20-Mc bandwidth over a distance of 1,000 km is calculated, assuming unfavorable conditions and basing the calculation on experimental results for cables (a) with central conductor of diameter 5 mm and outer conductor of internal diameter 18 mm, the insulation consisting of spiral ribbons of styroflex and the outer conductor formed of two half shells of copper, (b) of corresponding dimensions 2.6 and 9.4 mm, disks of of polythene being used for insulation. The mean-square values of phase and amplitude distortion are 0.05 μs and 1.6 per cent respectively for a 5/18-mm pair cable. The corresponding values for a 2.6/9.4 mm pair are significantly lower: 0.001 μs and 0.04 per cent.

621.315.212.017.71 28
Heating of Radio-Frequency Cables—W. W. Macalpine. (*Trans. AIEE*, vol. 68, Part I, pp. 283–288; 1949.) See 3014 of 1948.

621.315.213.12:621.315.221:621.3.011.4 29
Sheathed Lecher System—R. Gans. (*Z. Naturf.*, vol. 3a, pp. 519–521; 1948.) A formula is derived for the capacitance between the two wires of a Lecher system within a sheath. The degree of approximation is considerably better than obtained with Breisig's formula. A numerical example is worked. See also 2416 of 1950 (Wise).

621.392.211 30
The Propagation of Waves along an Endless Helix—S. Kh. Kogan. (*Compt. Rend. Acad. Sci.* (URSS), vol. 66, pp. 867–870; June 11, 1949. In Russian.) An equation (12) is derived determining the current distribution along the helix and methods for its solution are indicated. The theoretical results obtained have been confirmed experimentally.

621.396.67 31
Radiation from Circular Current Sheets—W. R. LePage, C. S. Roys, and S. Seely. (Proc. I.R.E., vol. 38, pp. 1069–1072; September, 1950.) A theoretical treatment of radiation from a system of current elements arranged on the circumferences of concentric circles. Series formulas are given for the radiation field produced by a specified excitation of these elements. Integral formulas are given for the excitation required to produce any desired radiation pattern. There is no consideration of how the required excitation can be produced.

621.396.67:538.566 32
Cylindrically Diverging Electromagnetic Waves in a Medium with Nonuniform Electrical Properties (Elias-Layer) above a Semiconducting Earth—C. T. F. van der Wyck. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 195–209; July and September, 1950.) Expressions in terms of the Hertzian vector potential are derived for the field of a vertical dipole above a flat, homogeneous, imperfectly conducting earth in a medium with refractive index depending exponentially on height. A geometrical-optical interpretation of the expressions is given, using the saddle-point method.

621.396.67:621.396.9 33
Radar Aerial Systems for Uniform Irradiation of a Surface—Huynen. (See 117.)

621.396.671 34
Impedance Transformation in Folded Dipoles—R. Guertler. (Proc. I.R.E., vol. 38, pp. 1042–1047; September, 1950.) Reprint. See 3342 of 1949.

621.396.671 35
Input Impedance of Horizontal Dipole Aerials at Low Heights above the Ground—R. F. Proctor. (*Proc. IEE*, Part III, vol. 97, p. 321; September, 1950.) Discussion on 2134 of 1950.

621.396.677 36
Microwave Lenses—J. Brown and S. S. D. Jones. (*Electronic Eng.* (London), vol. 22, pp. 127–131, 183–187, 227–231, 264–268, 358–362, and 429–434; April–July, September and October, 1950.) A comprehensive review of the subject, including theory of the principal types of lens.

621.396.677 37
Development of Artificial Microwave Optics in Germany—O. M. Stuetzer. (Proc. I.R.E., vol. 38, pp. 1053–1056; September, 1950.) Discussion of the development of microwave delay lenses using parallel metal strips or waveguide

sections. A lens of the latter type, for use on wavelengths of about 5 cm and with a diameter of 300 cm, is illustrated.

621.396.677 38
Factors Governing the Radiation Characteristics of Dielectric-Tube Aerials—D. G. Kiely. (*Proc. IEE*, Part III, vol. 97, pp. 311–321; September, 1950.) The effects of changes of tube length, diameter, and wall thickness on the radiation pattern were investigated experimentally. It is suggested that the mechanism of radiation of thin-walled dielectric tubes more closely resembles that of a lens than that of a leaking waveguide, such as a dielectric rod antenna. The gain of a dielectric-tube antenna of length 8 λ, diameter 1.16 λ, and wall thickness 0.03 λ is approximately 21 db.

621.396.677 39
Pattern Calculations for Antennas of Elliptical Aperture—R. J. Adams and K. S. Kelleher. (PROC. I.R.E., vol. 38, p. 1052; September, 1950.) The aperture illumination patterns in the directions of the major and minor axes are calculated from a knowledge of the patterns of the feed horn, and then expressed as Fourier series of up to four terms. The radiation pattern is given by

$$F(u_1, u_2)=\pi ab\sum\sum a_r+b_sG_{rs}(u_1, u_2),$$

where $u_1=(2\pi a \sin\phi)/\lambda, u_2=(2\pi b \sin\theta)/\lambda, a$ and b are the semimajor and semiminor axes of the aperture, a_r and b_s coefficients of the Fourier series and G_{rs} a complicated function which has been tabulated elsewhere by the authors. Very good agreement with the theory was obtained in experiments on several horns.

621.396.677 40
The Radiation of 'Beam' Aerials in Particular and of Large Surfaces in General—B. van der Pol. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 151–155; July and September, 1950.) The calculation of the power radiated by a beam antenna is simplified by imagining the system of parallel-wire radiators replaced by an equivalent extended current-carrying surface. Approximate formulas are derived for the radiated power for different limiting cases of the dimensions. Where the dimensions are large compared with λ the radiated power is independent of λ and is proportional to the area of the surface.

621.396.677 41
Measured Directivity Induced by a Conducting Cylinder of Arbitrary Length and Spacing Parallel to a Monopole Antenna—F. R. Abbott and C. R. Fisher. (PROC. I.R.E., vol. 38, pp. 1040–1041; September, 1950.) Curves are given for determining the directivity, given the separation of the parasite from the antenna and also its height.

621.396.679.4:621.392.094 42
Effects of Linear Distortion on a Band of Frequencies Transmitted along a Long Mismatched Line—J. Fagot. (*Ann. Radioélect.*, vol. 5, pp. 179–184; July, 1950.) The following formulas are derived for the irregularities caused by the mismatch when the frequency varies within the band considered:

period of variations $\Delta f=v/2e$

max./min. amplitude variation
$=(1+\rho\rho')/(\rho+\rho)$

phase displacement or propagation-time deviation $\Delta\tau=(\rho-1)(\rho'-1)e/v$

where v is the phase velocity, e the length of line, and ρ and ρ' the swr at the ends of the line. These formulas are independent of the carrier frequency and hold for the usual practical case where the terminal impedance is not highly selective.

621.396.679.4:621.392.094:621.396.619.13 43
Study of the Effects of a Long Line on a Frequency-Modulation Signal: Distortion, Compensation and Applications—M. Denis. (*Ann. Radioélect.*, vol. 5, pp. 185–205; July, 1950.) The long feeder lines used in cm-wave transmission are a source of distortion; this is particularly severe when the modulated oscillator is tightly coupled to a long, slightly mismatched feeder. The only effective remedy is the insertion of an amplifier between source and feeder. The existence of several junctions in a feeder, each causing slight reflection, may transform the line into a dispersive quadripole, so that phase distortion occurs, accompanied in certain cases by nonlinear frequency distortion. Numerical examples show the importance of this. Long lines are best avoided for high-quality transmissions using a cm-wave carrier with FM. Methods of correcting distortion and possible uses of long lines in measurement technique are discussed.

CIRCUITS AND CIRCUIT ELEMENTS

537.312.6:621.315.592 44
Semiconductors with Large Negative Temperature Coefficient: Thermistors—N'Guyen Thien-Chi and J. Suchet. (*Ann. Radioélect.*, vol. 5, pp. 155–167; July 1950.) Review of the development of the thermistor in the CSF laboratories, and discussion of properties and applications.

621.314.12:621.317.088.4 45
The Fundamental Limitations of the Second-Harmonic Type of Magnetic Modulator as Applied to the Amplification of Small D.C. Signals—Williams and Noble. (See 142.)

621.314.3† 46
Dynamoelectric Amplifiers—R. M. Saunders. (*Elec. Eng.*, vol. 69, pp. 711–716; August, 1950.) Basic principles are presented, five types are distinguished, salient features discussed, and methods proposed for predicting performance.

621.314.3† 47
Analytical Determination of Characteristics of Magnetic Amplifiers with Feedback—D. W. ver Planck, L. A. Finzi, and D. C. Beaumariage. (*Trans. AIEE*, vol. 68, Part I, pp. 565–570; 1949.) See also 2447, 2448 and 3064 of 1949 (ver Planck, Fishman, and Beaumariage).

621.314.3† 48
A New Theory of the Magnetic Amplifier—A. G. Milnes. (*Proc. IEE*, (London), Part II, vol. 97, pp. 460–474; Discussion, pp. 474–483; August, 1950.) Assuming that the B/H curve for the core material has a constant slope up to saturation level, followed by zero slope, flux waveforms are derived and equations developed for the magnetomotive forces operating throughout the cycle for a transductor with any degree of self excitation. Analytical expressions are thence derived for the output characteristics and for the current amplification and time constant of a transductor, which are of considerable importance in design work.

"An important phenomenon explained by the new theory is the variation of the current amplification with change of load resistance. This is readily detected in practice even for a simple transductor, and becomes of increasing significance as the percentage of self excitation is increased. The assumptions made in the analysis are such that the theory can be applied successfully only to transductors with cores which have high permeability and are readily saturable."

621.314.3† 49
New Core Materials Widen Scope of Saturable Reactors—. (*Elec. Mfg.*, vol. 42, pp. 126–214; September, 1948.) A general account of the subjects discussed in papers presented at a symposium on magnetic materials at the Naval Ordnance Laboratory, Washington, June 15, 1948.

621.316.8:621.396.822 50
Distribution in Energy of Johnson Noise Pulses—B. R. Gossick. (*Jour. Appl. Phys.*, vol. 21, pp. 847–850; September, 1950.) An analysis is made using Maxwell-Boltzmann statistics with a time-energy phase space. The results are checked against pulse-height measurements made with a linear amplifier and electronic counter. The following parameters are determined from the distribution function: (*a*) time of flight associated with a Lorentz mean free path; (*b*) number of electrons producing pulses of a given height; (*c*) the potential through which an electron falls along a Lorentz mean free path.

621.317.35:621.39.001.11 51
Signals with Limited Spectra and Their Transformations—J. Oswald. (*Câbles & Trans.* (Paris), vol. 4, pp. 197–215; July, 1950.) Detailed mathematical treatment. See also 3069 of 1949.

621.318.371:621.392.53 52
The Compensation of Delay Distortion in Video Delay Lines—R. A. Erickson and H. Sommer. (PROC. I.R.E., vol. 38, pp. 1036–1040; September, 1950.) A theoretical and practical investigation into the effect of placing isolated metal patches of various sizes in the vicinity of solenoid delay lines. These patches are equivalent to capacitors bridging sections of the solenoid. The relation between size of patch and effective capacitance is given. Relations necessary for determining the delay distortion and effective bandwidth corresponding to different amounts of capacitance are determined. Oscillograms illustrate the reduction of distortion due to patches of different sizes. See also 41 of 1947 (Kallmann).

621.319.4 53
Metallized Paper Capacitors—J. R. Weeks. (PROC. I.R.E., vol. 38, pp. 1015–1018; September, 1950.) See also 134 below (McLean).

621.392 54
Simplifying Assumptions—W. T. C. (*Wireless Eng.*, vol. 27, pp. 217–219; August and September, 1950.) Examples are discussed which illustrate the need for great care in making simplifying assumptions. Neglect of the leakage inductance of a transformer, for instance, results in considerable discrepancy between the wave form to be expected in a branch of a television line-scanning circuit and the wave form actually observed.

621.392 55
The Distinction between Effective and Circuit Bandwidths—W. J. Kessler. (*Trans. AIEE*, vol. 68, Part I, pp. 98–99; 1949.) See 3074 of 1949.

621.392 56
Wideband Two-Phase Networks—N. O. Johannesson. (*Wireless Eng.*, vol. 27, pp. 237–238; August and September, 1950.) Comment on 1356 of 1950. A method of approximation based on θ functions is given which facilitates numerical computations.

621.392.5 57
Application of the Methods of Synthesis of Electrical Circuits to the Construction of Quadripoles, Given the Transmission Coefficient for a Finite Frequency Band—V. A. Taft. (*Bull. Acad. Sci.* (URSS), pp. 873–887; June, 1950. In Russian.) The methods used for designing a quadripole are normally based on trials of different values for circuit parameters. A general design method based on the synthesis of electrical circuits is here proposed. The theory of quadripoles is discussed and the necessary and sufficient conditions are established for the physical realization of a passive reactive quadripole. The various stages of the design work are then laid out for realizing a quadripole with a given transmission coefficient

for a finite frequency band or, in other words, with given attenuation and phase-displacement coefficients for the same frequency band.

621.392 58
Summary of Transformations Useful in Constructing Analogs of Linear Vibration Problems—J. P. Corbett. (*Trans. AIEE*, vol. 68, Part I, pp. 661–664; 1949.) Full paper. Summary noted in 834 of 1950.

621.392:519.272.15 59
Short-Time Autocorrelation Functions and Power Spectra—Fano. (See 141.)

621.392:681.142 60
A High-Speed Multiplier for Analogue Computers—B. N. Locanthi. (*Elec. Eng.*, vol. 69, p. 717; August, 1950.) Summary of AIEE Summer General Meeting paper. A circuit developed at the California Institute of Technology for multiplying together two voltages is described. One voltage is fed to a ring modulator; the output, together with the other voltage is fed to a balanced modulator whose output provides the required product. Phase and amplitude distortion are low.

621.392.4 61
Two-Pole Compensation Networks—A. Pinciroli. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 644–646; August 20, 1949. In French.) Paper presented at the International Television Conference. Zürich, 1948. Discussion of the characteristics of a triode circuit equivalent to a tube with negative transconductance. Such a circuit, when connected between two points of an electrical network, can within certain limits compensate the resistive and capacitive parameters of the network between the two points. Applied to the output terminals of a *RC* amplifier, a compensation two-pole can increase the upper frequency limit considerably and also the amplification factor. See also 1882 of 1948.

621.392.52:621.396.611.3 62
Tuned Absorption Circuits—R. E. Spencer. (*Wireless Eng.*, vol. 27, pp. 219–224; August and September, 1950.) An analysis is given of circuits in which input and output terminals are connected to the same tuned circuit and a coupled circuit absorbs power over a comparatively narrow band of frequencies. Approximate mathematical solutions are given, with graphical illustrations of response curves and a discussion of the depth and shape of the trough between the two humps of the response curve.

621.392.52:621.396.619 63
Polyphase Modulation as a Solution of Certain Filtration Problems in Telecommunication—Macdiarmid and Tucker. (See 213.)

621.392.6 64
Synthesis of 2*n*-Terminal Passive Networks—R. Leroy. (*Câbles & Trans.*, (Paris), vol. 4, pp. 234–247; July, 1950.) Analytical investigation of certain aspects of the theory discussed in 62 of 1950 (Bayard). By extension of Gewertz's method to a positive real completely reduced matrix of order n it is possible to obtain a matrix of the same order but with elements of degree lower by $(2n-2)$. A singular type of matrix appears in the calculations when the degree of the elements is $<(2n-2)$, so that the reduction of the initial matrix generally involves the derivation from a singular matrix of a non-singular matrix of lower order, to which the Gewertz method is then applied. The corresponding network always includes only one resistor. An alternative method for effecting the synthesis of completely reduced matrices consists in determining the reactive network which includes this resistor between a supplementary pair of terminals. Further, reduced positive real matrices of row q can be realized as networks comprising q resistors.

621.392.6 65
Synthesis of a Finite 2*n*-Terminal Network by a Group of Networks each of which Contains Only One Ohmic Resistance—Y. Oono. (*Jour. Math. Phys.*, vol. 29, pp. 13–26; April, 1950.)

621.396.611.1 66
Periodic and Aperiodic Oscillations in an Oscillatory Circuit incluidng an Iron-Cored Coil and Approximately Tuned to an Impressed Alternating Voltage—J. M. Op den Orth. (*Tijdschr, ned. Radiogenoot.*, vol. 15, pp. 211–236; July and September, 1950.) The investigation relates particularly to a circuit including little or no resistance. The inductance is assumed to be only slightly nonlinear, thus permitting the circuit differential equations to be solved by an approximate method, due to Poincaré. Periodic forced oscillations are represented by singular points (generally two isolated points and a saddle-point). If damping is zero, all other solutions correspond to closed curves around one or the other of the isolated points, or around both points. If damping is not zero the curves become spirals tending to one or the other of the isolated points (corresponding to stable forced oscillations). Jump phenomena are discussed. The importance of the solution curves through the third singular point (a saddle-point representing an unstable forced oscillation) is made clear.

621.396.611.1 67
Nonharmonic Oscillations as Caused by Magnetic Saturation—R. Rüdenberg. (*Trans. AIEE*, vol. 68, Part I, pp. 676–685; 1949.) A detailed investigation of the effect produced on the waveform by the nonlinear characteristics of circuits containing capacitors and saturated inductors. Rigorous analysis shows that the natural oscillations in such circuits remain harmonic and have constant frequency only at small amplitudes. With increasing amplitude the wave form becomes more and more distorted and the natural frequency increases. Transient and steady-state forced oscillations are treated by means of a differential equation which involves only four parameters, but which covers all possible cases. A solution is obtained by a step-by-step method. In the transient state, strange shapes are found for the flux, current and voltage curves, with no regularity as regards either symmetry or periodic repetition. Natural oscillations can be maintained by application of a voltage of definite magnitude and wave form. In series circuits a peaked voltage curve is required. Subharmonics of the supply voltage may consequently be developed. The theoretical results are illustrated by oscillograms obtained with circuits including highly saturated components.

621.396.611.1 68
The Fourier Spectrum of Forced Oscillations Produced by Step and Needle Impulses—M. Päsler and W. Reichardt. (*Frequenz*, vol. 4, pp. 211–215; August, 1950.) The combined effect of a voltage step and its derivative on an oscillatory circuit is analysed by means of Fourier integrals, and the dependence of the spectral functions on frequency and damping is investigated. The results are presented in graphs and discussed. See also 1324 of 1949 (Päsler).

621.396.611.21 69
High-Frequency Vibrations of Plates Made from Isometric and Tetragonal Crystals—E. A. Gerber. (Proc. I.R.E., vol. 38, pp. 1073–1078; September, 1950.) Bevelling of crystals is described as a method for obtaining a single response frequency in crystal units. Electrical characteristics and temperature coefficients of $NaClO_3$, $NaBrO_3$, $HN_4H_2PO_4$(ADP) and KH_2PO_4(KDP) crystal units were measured. The two cuts (zxw) 45° and ($zxtw$) 45°/54°44′ [See IRE Standard on Piezoelectric Crystals (655 of 1950)] were used. Fair agreement was obtained with the theory presented in the paper. The $NaBrO_3$ thickness modes have about the same quality factor as that of quartz, the quality factor of ADP crystals being about one order of magnitude lower.

621.396.611.4 70
Some Perturbation Effects in Cavity Resonators—A. Cunliffe and L. E. S. Mathias. (*Proc. IEE* (London), Part III, vol. 97, pp. 367–376; September, 1950.) "An investigation, partly theoretical and partly experimental, has been made of some effects which occur when the boundary of a cavity is deformed slightly. First, the theory of natural electromagnetic oscillations inside lossless cavities is summarized. Then a general theory, following along the lines of conventional first-order perturbation theory, is given. The theory has been applied to the perturbation of a right-circular cylindrical cavity. Two cases have been considered: the E_{010} mode, a non-degenerate case, and the H_{011} and E_{111} modes, a triply degenerate case. These two cases have also been investigated experimentally. Theory and experiment are in reasonable agreement, even for quite large deformations, when the deformation is applied gradually over a large area of the cavity wall. For sharp abrupt changes in the geometry of the cavity wall, however, it appears that the first-order perturbation theory can be applied only for very small distortions. The general results, theoretical and experimental, which have been obtained, show that if the frequency of the operating mode is well separated from the frequencies of other modes, a deformation of the boundary changes only the frequency of the operating mode and not its electromagnetic field configuration. If the frequency of the operating mode is near to the frequencies of other modes, a slight deformation of the cavity boundary, as well as changing the frequency of the operating mode, may also change its electromagnetic field configuration. 'Lossy' material or resistance wires, introduced into a cavity with a view to damping out unwanted modes, may also affect the desired resonance if certain types of deformation are present."

621.396.615 71
The Reactance-Tube Oscillator—A. Giger. (Proc. I.R.E., vol. 38, p. 1096; September, 1950.) Comment on 326 of 1950 (Chang and Rideout), pointing out that the transconductance of the oscillator tube has a fixed value and hence cannot affect the frequency. An explanation of the observed frequency variation is given which is based on practical work in Switzerland, where the reactance tube oscillator has been in commercial use for years.

621.396.615.17 72
Calculation of the Time Delay of a Multivibrator—H. de Lange Dzn. (*Tjdschr. ned. Radiogenoot.*, vol. 15, pp. 275–291; July and September, 1950.) A general solution is derived for a linear differential equation of nth order with constant coefficients by the method of variation of parameters, in which a discontinuity of the nth order is reduced to one of zero order during integration. The solution is applied to calculate the time delay of a multivibrator by means of a differential equation of fourth order.

621.396.615.17:621.317.755 73
Spiral Time Base—(*Wireless Eng.*, vol. 27, pp. 224–226; August and September, 1950.) A description of the mode of operation of the basic circuit, with particular attention to the effect of varying the time at which the trigger switch is opened. A simple circuit using a thyratron switch is shown, and a more complex switching circuit used in radar is fully described. The timebase was developed at the Radar Research and Development Establish-

ment and is the subject of British Patent No. 582419.

621.396.645 74
A Selective Relay Amplifier for Recording WWV Time Signals—E. F. Carome and H. C. Nash. (*Trans. Amer. Geophys. Union*, vol. 30, vol. 30, pp. 328–329; June, 1949.) Full circuit details are given of an amplifier which discriminates against all frequencies except 440 cps, the modulation frequency of the wwv signals, so that the operation of a relay in the anode circuit of the output tube is unaffected by the pulses sent out at second intervals, or by atmospherics or background noises. The selective element consists of a twin-T *RC* bridge.

621.396.645:612.8 75
Biological Requirements for the Design of Amplifiers—H. Grundfest. (Proc. I.R.E., vol. 38, pp. 1018–1028; September, 1950.) Discussion of the nature and properties of bioelectric potentials and description of amplifiers specially suitable for investigating such potentials.

621.396.645:621.3.015.7† 76
The Amplification of Pulse-Form Modulated Voltages and the Accompanying Reduction of Slope and Time Delay—J. W. Alexander. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 237–251; July and September, 1950.) In assessing the performance of an amplifier dealing with pulse signals the criteria to be considered are the slope of the output-pulse envelope and its retardation with respect to the input pulse. According to the method developed, these factors are easier to determine than the output-pulse envelope itself. The theory is applied to amplifiers with several single-tuned circuits and to amplifiers with coupled–and stagger-tuned circuits.

621.396.645.012.3 77
The Determination of Quiescent Voltages and Currents in Pentode Amplifiers—A. J. Shimmins. (*Electronic Eng.*, vol. 22, pp. 386–388; September, 1950.) A method is described for determining the quiescent values of anode, screen-grid and control-grid voltages and currents graphically or by calculation from a family of dynamic characteristics for various screen voltages. The solution is approximate but is useful for predicting effect due to changes in circuit parameters such as cathode and screen resistances.

621.396.645.36 78
"The Cathamplifier"—C. A. Parry. (Proc. I.R.E. (Australia), vol. 11, pp. 199–204; August, 1950.) An amplifier circuit with high input impedance which permits push-pull operation from an unbalanced source. The input voltage is applied between earth and the grid of one of the tubes, and a voltage proportional to the total circulating current is obtained from a transformer whose center-tapped primary is connected between the two cathodes; this voltage is applied, in the correct phase, between earth and the grid of the other tube. A resistance shunted across the transformer primary is varied to obtain anode-ac balance. Over-all performance similar to that usual in push-pull operation may be obtained. Two modes of oscillation are possible which are independently adjustable.

GENERAL PHYSICS

533.723+621.396.822 79
Spontaneous Fluctuations—D. K. C. MacDonald. (*Rep. Progr. Phys.*, vol. 12, pp. 56–79. References, pp. 79–81; 1948–49.) A survey of developments of fluctuations analysis, and a review of research on fluctuation phenomena in the last decade, omitting general problems treated in standard works. The correlation function is discussed, with examples of its use. Recent developments treated are mainly concerned with electrical and tube noise, and include discussion of thermal, shot, and low-frequency noise.

535.215.9:537.315 80
Contact-Potential Measurements on Irradiated Metal-Oxide Surfaces—H. Neuert. (*Z. Naturf.*, vol. 3a, pp. 226–228; 1948.) Account of experimental work demonstrating that a slightly oxidized surface can be activated by sw ultraviolet radiation, or by ionic or electronic charging, or by mechanical treatment.

535.312:539.23 81
Reflection Reduction in Optics—M. Auwärter. (*Bull. schweiz. elektrotech, Ver.*, vol. 40, pp. 605–607; August 20, 1949. In German.) Paper presented at the International Television Conference, Zürich, 1948. Investigation of the properties of single- and multiple-surface layers.

535.317.9:621.397.5 82
The Schmidt Optical System—H. Rinia. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 580–585; August 20, 1949. In English.) Paper presented at the International Television Conference, Zürich, 1948. An outline is given of the basic principles of the Schmidt system. The conventional form gives fifth-order coma for low magnifications. Means are indicated for compensating this coma. The cause and magnitude of the lateral spherical aberration are discussed and a new method of aberration correction is described. See also 63 and 1215 of 1949 (Rinia and van Alphen).

535.42 83
Critical Report on the General Laws of Diffraction, Submitted to the International Optics Commission—G. Toraldo di Francia. (*Nuovo Cim.*, vol. 5, pp. 591–605; December 1, 1948.) A review of classical and modern theories.

537.312.5 84
Theory of Photoelectric Conduction in Composite Conductors—F. Stöckmann. (*Z. Phys.*, vol. 128, pp. 185–211; July, 1950.) The fundamental equations of electrical conduction are examined for the case of circuits including electron sources and sinks. General laws for photoelectric currents are hence derived which agree with laws derived directly by other workers, and saturation current, amplification and exponential law of loss are discussed. Nonlinear conductors and semiconductors are studied as special cases.

537.525 85
The High-Frequency Gas Discharge—F. Kirchner. (*Z. Naturf.*, vol. 2a, pp. 620–621; 1948.) A description is given of a simple experimental arrangement for investigating the discharge without introducing a probe. The results indicate strong positive space-charge and high positive potential within the gas. The discharge can be maintained with voltages below the ionization potential.

537.528 86
X-Ray Investigation of Sound Waves Associated with Breakdown of Dielectrica—W. Schaaffs and F. Trendelenburg. (*Z. Naturf.*, vol. 3a, pp. 656–668; 1948.)

537.533 87
General Properties of the Electronoptical Image—F. Borgnis (*Bull. schweiz. elektrotech. Ver.*, vol. 40, p. 571; August 4, 1949. In German.) Paper presented at the International Television Conference, Zürich, 1948. See also 1351 and 2780 of 1949, for which the above UDC number is preferable.

537.534:621.385.82 88
The Production of Ion Beams by Means of a High-Frequency Discharge—H. Neuert. (*Z. Naturf.*, vol. 3a, pp. 310–312; 1948.) An arrangement similar to that described by Thonemann (3261 of 1946) was studied. With 100 w exciting power and 10 kv field voltage an ion current of 20 ma was obtained.

538.3 89
A Notation for Electrodynamics Adaptable to any System of Units—R. Fleischmann. (*Z. Naturf.*, vol. 3a, pp. 492–495; 1948.) The fundamental equations of electricity and the principal formulas of four-dimensional electrodynamics are presented in a notation which is independent of the system of units and which for special cases yields the formulas valid for the usual systems of units.

538.31+538.65 90
Fields within and around Cavities in a Magnetically Strained Medium, Ponderomotive Forces acting thereon in a Magnetic Field with Current in the Cavity, and the Electric Field generated on Movement of the Cavity—J. P. Schouten. (*Tijdschr, ned. Radiogenoot.*, vol. 15, pp. 163–177; July and September, 1950.) It is shown analytically that the total electric field inside a cavity moving in a medium subjected to a homogeneous magnetic field is independent of the inhomogeneity introduced by the cavity wall, and the total mechanical force on the cavity is the same as if the medium were continuous.

538.311 91
Formulas and Tables for the Calculation of the Magnetic Field Components of Circular Filaments and Solenoids—F. W. Grover. (*Trans. AIEE*, vol. 68, Part I, pp. 665–675; 1949.) Existing formulas are discussed and new formulas are derived. Tables of the functions involved are presented in a form which facilitates routine calculations.

538.322 92
The Forces between Two Current Conductors—B. D. H. Tellegen. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 157–161; July and September, 1950.) Conditions for steady currents in closed conductors are considered, and the total force is regarded as the resultant of attractive forces and couples due to the interaction of current elements of the two conductors.

538.56:535.3 93
Reflection and Transmission of Electromagnetic Waves by Thin Curved Shells—J. B. Keller. (*Jour. Appl. Phys.*, vol. 21, pp. 896–901; September, 1950.) "The scattering of an arbitrary electromagnetic field by a conducting or non-conducting obstacle is investigated. The differential equations and boundary conditions satisfied by the field are transformed into a pair of inhomogeneous linear integro-differential equations for E and H. For an obstacle which is a thin shell of constant thickness h, a formal procedure for obtaining a solution of these equations as power series in h is given. The lowest order term in this solution is the incident field. An explicit expression for the next term is found in the form of a surface integral. This integral is evaluated approximately by the method of stationary phase. The physical properties of the solution are examined in detail, and satisfactory agreement are examined in detail, and satisfactory agreement is found with many results previously obtained by other methods."

538.56:535.42 94
Rigorous Theory of the Diffraction of Electromagnetic Waves by a Perfectly Conducting Disk—J. Meixner. (*Z. Naturf.*, vol. 3a, pp. 506–518; 1948.) The solution of this problem is given and is extended to the related problem of diffraction by a circular aperture in a perfectly conducting plane sheet of infinite extent by a generalization of Babinet's principle.

538.56:537.56:523.74 95

The Characteristics of Radio-Frequency Radiation in an Ionized Gas, with Applications to the Transfer of Radiation in the Solar Atmosphere—S. F. Smerd and K. C. Westfold. (*Phil. Mag.*, vol. 40, pp. 831–848; August, 1949.) The function E which determines the radiation at any point is the ratio of the emissivity to the product of the absorption coefficient and the square of the index of refraction. The intensity of 'quiet' radio-frequency solar radiation reaching the earth can be expressed in terms of E and the optical depth of the various ray trajectories. Formulas are derived for emissivity, absorption coefficient and refractive index, from which E can be found. The formulas are expressed in terms of the electron and ion densities and the kinetic temperature, assuming a Maxwellian velocity distribution. A heuristic theory of the absorption and emission processes on a volume element is given which takes account of the effect of the surrounding particles of the medium.

538.56.029.64:537.562:535.43 96

The Scattering of 3-cm Radiation by Ionized Gases—S. N. Denno, H. A. Prime, and J. D. Craggs. (*Proc. Phys. Soc.*, vol. 63, pp. 726–727; September 1, 1950.) Radiation of wavelength 3 cm is scattered by a commercial cylindrical mercury discharge tube. Scatter at right angles to the incident radiation is received and its power measured. Curves showing the received power as a function of tube current are given for radiation polarized (*a*) perpendicular to and (b) parallel to the tube axis, for two tubes of diameters 3.1 cm and 6.4 cm respectively. It is intended to use the information to determine the electron concentration in the discharge.

539 97

Masses of Fundamental Particles—R. Fürth. (*Nature* (London), vol. 166, pp. 727–728; October 28, 1950.)

548.1:537 98

A Derivation and Tabulation of the Piezoelectric Equations of State—J. F. Haskins and J. S. Hickman. (*Jour. Acoust. Soc. Amer.*, vol. 22, pp. 584–585; September, 1950.) The conservation-of-energy principle is applied to derive the general equations, with strain, electric displacement and entropy as independent variables. The special case of constant entropy is then considered, with polarization as an additional parameter, and all possible linear adiabatic equations of state are developed, using in turn as independent variables stress and electric field, strain and polarization, and stress and polarization. Hence the relations between the elastic, electric and piezoelectric coefficients for the various pairs of independent parameters are determined and tabulated.

548.1:537 99

Piezoelectric Equations of State and their Application to Thickness-Vibration Transducers—W. G. Cady. (*Jour. Acoust. Soc. Amer.*, vol. 22, pp. 579–583; September, 1950.) The electromechanical equations of state are given in several forms and those most appropriate in theoretical work are indicated. A detailed treatment of the thickness-vibration transducer is given, resulting in expressions for the electrical characteristics and acoustic power. Various special cases are briefly considered.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:[621.396.9+77 100

Meteor Velocities—P. M. Millman and D. W. R. McKinley. (*Observatory*, vol. 70, pp. 156–158; August, 1950.) Photographic and radio techniques are discussed in relation to their capabilities of detecting fast meteors. Meteors with speeds as high as 150 km should be detectable by radio methods.

523.7:538.12 101

The Effect of Turbulence on a Magnetic Field—P. A. Sweet. (*Mon. Not. R. Astr. Soc.* vol. 110, pp. 69–83; 1950.) Extension of earlier work (1677 of 1950). Mathematical analysis indicates that (*a*) turbulence reduces the effective conductivity in the core and in the outer layers of the sun, but sunspot fields are not affected; (*b*) the decay time of the sun's general magnetic field is somewhat less than 10^{10} years, while the mean field in the core, if the general field is decaying from some initial state, is irrotational; (*c*) no system of meridian-plane convection currents in the sun can provide a general amplification of a field produced by a given emf, while turbulence in fact reduces the field.

523.71 102

The Solar Constant—C. W. Allen. (*Observatory*, vol. 70, pp. 154–155; August, 1950.) Discussion leads to a tentative value of 1.97 cal/cm^2/min.

523.72 103

Cosmic Radiation from the Sun—A. Ehmert. (*Z. Naturf.*, vol. 3a, pp. 264–285; 1948.) The phenomena of chromosphere eruptions may be related to the ionization of chromosphere regions by protons accelerated in the varying magnetic fields of sunspots; in the same way accelerated electrons may give up their energy as usw radiation.

523.72:621.396.822 104

Solar Radio-Frequency Radiation—J. L. Pawsey. (*Proc. IEE* (London) Part III, vol. 97, pp. 290–308; September, 1950. Discussion, pp. 308–310.) A survey of solar-noise research from 1932 to 1948. Observed characteristics in the wavelength range from 1 cm to a few meters are described. From the intensity, region of origin, association with visual phenomena and polarization a classification is suggested. A thermal component is recognized corresponding in intensity to a black-body radiation of a temperature that rises from $10^{4\circ}$K at a wavelength of 1 cm to $10^{6\circ}$K at wavelengths of a few meters. This is believed to be associated with a rise in the region of origin from the lower chromosphere to the corona. Non-thermal components, prominent at meter wavelengths and believed to originate from electrical disturbances in the solar atmosphere, vary rapidly and have occasional peak intensities 10^3 to 10^6 times the thermal ones. 66 references are given.

523.74:538.56:537.56 105

The Characteristics of Radio-Frequency Radiation in an Ionized Gas, with Applications to the Transfer of Radiation in the Solar Atmosphere—Smerd and Westfold. (See 95.)

523.74/.75:621.396.11 106

Solar Notes—H. W. Newton. (*Observatory*, vol. 70, pp. 163–164; August, 1950.) Solar activity during the first six months of 1950 is briefly reviewed. Observed correlations between special solar phenomena, such as flares, and radio propagation conditions are described.

523.8:538.12 107

Stellar Magnetic Fields and Rotation—S. K. Runcorn. (*Observatory*, vol. 70, pp. 155–156; August, 1950.) Using data applying to eight stars, the correlation coefficient between magnetic field and angular momentum is 0.6. Possible causes of varying and reversing magnetic fields are discussed.

538.12:521.12 108

Gravitational Field and Magnetism—A. Maior. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231, pp. 607–608; September 25, 1950.) Argument indicating that a charged body moving in a gravitational field produces a convection current accompanied by a magnetic field. Formulas are derived which are more general than the empirical formula of Blackett.

550.372 109

The Earth's Constants from Combined Electric and Magnetic Measurements partly in the Vicinity of the Emitter—K. F. Niessen. (*Z. Naturf.*, vol. 3a pp. 552–558; 1948. In English.) The dielectric constant and conductivity of the earth in the vicinity of a projected transmitter are found by making measurements of E and H at points close to (distance 2λ–4λ) and remote from an experimental transmitter. The method depends on the fact that at short distances the electric and magnetic field strengths vary according to different laws. The analysis is based on Sommerfeld's vector-function formula for the radiation from an ideal dipole, and the required constants are obtained from a simple system of curves. See also 3893 of 1947.

550.38 110

The 'Absolute Quadrupole-Moment'—a Fundamental Magnetostatic Quantity and its Geophysical Significance—H. G. Macht. (*Z. Naturf.*, vol. 3a, pp. 189–195; 1948.)

551.5:621.396.81.029.63/.64 111

A Radio Meteorological Investigation in the South Island of New Zealand—Milnes and Unwin. (See 200.)

551.510.535 112

Ionosphere Observations in Adélie Land—M. Barré and K. Rawer. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231, pp. 436–437; August, 16, 1950.) An interim report of observations made during the antarctic cruise of the *Commandant Charcot* (see also 3429 of 1949 and 97 of 1950). Features noted from recordings obtained during 24 hours on January 3–4, 1950 include: (*a*) a sporadic-E layer giving echoes at high frequencies (10 Mc); frequent stratification of this layer and increase of its height with frequency; (*b*) permanent diffusion of the F layer, extending with increase of frequency; (*c*) horizontal traces resembling those of the sporadic-E layer but at heights of 250–800 km; (*d*) selective absorption at about 3.5 Mc masking all trace of layers.

551.510.535:538.566 113

The Poynting Vector in the Ionosphere—Scott. (See 193.)

551.594.6(729) 114

Air Weather Service Sferics Operations in the Caribbean Area—H. F. Willey. (*Trans. Amer. Geophys. Union*, vol. 30, pp. 330–332; June, 1949.) A general description is given of the spherics 4-station network and its operation, with discussion of the correlation of spherics with weather conditions.

523.72+523.854]:621.396.822 115

Bruits Radio-Électriques Solaires et Galactiques (Solar and Galactic Radio Noise) [Book Review]—URSI (Union Radio Scientifique Internationale) Special Report No. 1. Brussels, 1950, 51 pp. (*HF, Brussels*, p. 200; 1950.) A résumé of present knowledge. The report was presented at the Stockholm meeting, 1948. An English edition is also available.

LOCATION AND AIDS TO NAVIGATION

534.88 116

Sound Ranging at the Morris Dam Torpedo Ranges—R. N. Skeeters. (*Elec. Eng.*, vol. 69, p. 718; August, 1950.) Summary of AIEE Summer General Meeting paper. For making accurate determinations of the trajectories of under-water missiles, the latter are caused to generate sounds which are picked up by suitably located hydrophones arrayed in groups of eight. An oscillograph record is obtained. Re-

duction of the data is accomplished by means of a computer which is a scale model of one of the hydrophone groups.

621.396.9:621.396.67 117

Radar Aerial Systems for Uniform Irradiation of a Surface—J. R. Huynen. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 293–297; July and September, 1950.) The problem of achieving uniformity of energy at all points of a spiral scan is studied.

621.396.93 118

The Relative Merits of Presentation of Bearings by Aural-Null and Twin-Channel Cathod-Ray Direction-Finders—S. de Walden and J. C. Swallow. (*Proc. IEE* (London), Part III, vol. 97, pp. 362–365; September, 1950.) Discussion on 3142 of 1949.

621.396.93 119

Some Experiments on the Accuracy of Bearings taken on an Aural-Null Direction-Finder—F. Horner. (*Proc. IEE* (London), Part III, vol. 97, pp. 359–361; Discussion, pp. 362–365; September, 1950.) The paper describes some tests to determine how the accuracy of a bearing taken on an aural-null rotating H-Adcock direction-finder depends on the width of the minimum and on the receiver output noise-level. The results pertain to bearings taken on a steady tone-modulated signal by an experienced observer working under good conditions. They indicate that bearings taken under these conditions will have standard deviation of between 1/40 and 1/20 of the arc of silence except for very small arcs, even when the bearings are derived from very few oscillations of the antenna system. Accuracy is improved if the number of complete oscillations is greater than about five. Accuracy is degraded if the angle through which the antenna is swung is increased to improve the quality of the signal at the limits of the swing.

Differences of at least two to one in the standard deviation of observed bearings may occur between different observers, and with one observer at different times. Compared with these changes, any changes due to the use of different receiver output noise-levels are considered to be small.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.787.9 120

The Pirani Effect in a Thermionic Filament as a Means of Measuring Low Pressures—(*Brit. Jour. Appl. Phys.*, vol. 1, p. 240; September, 1950.) Correction to paper abstracted in 1922 of 1950. W. P. Jolly is the sole author.

533.56 121

Characteristics of Diffusion Pumps—R. Witty. (*Brit. Jour. Appl. Phys.*, vol. 1, pp. 232–237; September, 1950.)

533.583:621.385 122

A Method for Measuring the Efficiency of Getters at Low Pressures—S. Wagener. (*Brit. Jour. Appl. Phys.*, vol. 1, pp. 225–231; September, 1950.) The method is based on measurement of the pressure drop along a narrow tube connecting the bulb containing the getter to the manifold of the pumping system. The values found for the rate of absorption of air ranged from 10 cm^3/s for Mg to 1,500 cm^3/s for Th getter, the bulb pressure being 1.5×10^{-6} mm Hg.

535.37 123

Phosphors and Phosphorescence—G. F. J. Garlick. (*Rep. Progr. Phys.*, vol. 12, pp. 34–53; References, pp. 53–55; 1948–49.) Recent investigations of luminescence in crystalline impurity-activated phosphors are reviewed. The electron-energy-band model and experimental support for it are discussed. Long-duration phosphorescence, due to electrons trapped in thermally metastable levels, correlates with thermoluminescence. Advances are reported in knowledge of the structure of luminescence emission centers in sulphide and silicate phosphors. The emission spectra of manganese-activated silicates, recent studies of oxides and tungstates, and infrared-sensitive phosphors are treated. The latter need a secondary activator for marked sensitivity; infrared light appears to cause the ejection of trapped electrons, but there is no simple correlation between optical and thermal ejection.

535.37:546.472.21 124

The Introduction of Copper into a Luminescent Zinc Sulphide—N. Ril' and G. Ortman. (*Compt. Rend. Sci.* (URSS), vol. 66, pp. 841–845; June 11, 1949. In Russian). If copper is introduced by diffusion into ZnS crystals it may be in two states, one causing blue luminescence and the other green luminescence. An experimental study of these effects is described.

537.311.31:546.92:541.183.56 125

Resistance Variation of a Platinum Foil due to Gas Adsorption—W. Braunbek. (*Z. Naturf.*, Vol. 3a, pp. 216–220; 1948.) Experiments made with Pt foil in oxygen, argon and helium indicate a resistance reduction of the order of 10^{-4} as compared with the value in vacuo, the effect being greatest in oxygen.

538.221 126

On Ferromagnetic States—J. Giltay. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 253–274; July and September, 1950.) The form of Madelung's laws is criticized and new expressions are formulated covering effects observed in ferromagnetic materials. Operating-cycle diagrams derived from auxiliary loop curves are introduced.

538.221:538.562 127

Magnetostriction of Permanent-Magnet Alloys—E. A. Nesbitt. (*Jour. Appl. Phys.*, vol. 21, pp. 879–889; September, 1950.) Magnetostriction measurements were made on various alloys having coercive forces from 50 to 600 oersted. In the older carbon-hardening permanent magnets, high coercive force and high magnetostriction appear together; for the newer carbon-free type this coincidence does not hold. These results are discussed in the light of recent theories.

548.0:537.228.1 128

Determination of the Elastic and Piezoelectric Coefficients of Monoclinic Crystals, with particular Reference to Ethylene Diamine Tartrate—R. Bechmann. (*Proc. Phys. Soc.*, vol. 63, pp. 577–589; August 1, 1950.) Longitudinal modes of vibration were used for narrow bars, low-frequency longitudinal and face-shear iodes for square plates containing the axis of symmetry, coupled modes for square plates perpendicular to the axis of symmetry, and thickness-shear modes for plates containing the axis of symmetry. New values of the coefficients and their temperature coefficients are given for EDT. Some properties are considered of face-shear vibrating square plates of EDT rotated about the axis of symmetry as functions of the orientation, and of Y-cut plates as functions of the width-to-length ratio.

548.0:549.451 129

Influence of Plastic Flow on the Electrical and Photographic Properties of the Alkali-Halide Crystals—F. Seitz. (*Phys. Rev.*, vol. 80, pp. 239–243; October 15, 1950.)

620.193:621.315.61 130

Methods for Determining the Effect of Contaminants on Electrical Insulation—K. N. Mathes, L. E. Sieffert, H. P. Walker, and R. H. Lindsey. (*Trans. AIEE*, vol. 68, Part I, pp. 113–118. Discussion, pp. 118–119; 1949.) The results of tests made under laboratory conditions with carefully controlled mixtures of such contaminants as are commonly encountered on board ship are tabulated according to the effects on the physical and electrical properties of insulating materials, including surface breakdown voltage, dimensional stability, and so forth.

621.315.61:621.317.331 131

Some Measurements of the Resistivity of Good Insulators—N. W. Ramsey. (*Proc. Phys. Soc.*, vol. 63, pp. 590–594; August 1, 1950.) A method depending on the loss of charge on a capacitor was used. The resistances of amber, alkathene, distrene and perspex increased over a period of weeks, the final values being considerably higher than previously published figures.

621.315.612.011.5 132

Ceramic Dielectrics with High Permittivity Titanates—A. Danzin. (*Ann. Radioélect.*, vol. 5, pp. 230–242; July, 1950. *Onde élect.*, vol. 30, pp. 253–258 and 335–340; June and July, 1950.) The mode of preparation of ceramic dielectrics is described and the properties of normal mineral insulating materials and of titanates are compared. Crystal structure and anomalous temperature coefficients are discussed and methods of obtaining a specified dielectric constant are outlined. Titanates are broadly classified into two groups and applications are listed.

621.315.612.4:621.3.011.5 133

The Structure, Electrical Properties and Potential Applications of the Barium-Titanate Class of Ceramic Materials—W. Jackson. (*Proc. IEE* (London), Part III, vol. 97, pp. 285–289; September, 1950.) Abstract of IEE lecture, March, 1950, reviewing present knowledge. The 75 references given are intended to afford a preliminary guide to the published work on the subject.

621.319.4:[621.793:621.315.614.6 134

Metallized Paper for Capacitors—D. A. McLean. (PROC. I.R.E., vol. 38, pp. 1010–1014; September, 1950.) An account of development work in the Bell laboratories. Lacquering the paper prior to metallizing increases the dielectric strength and insulation resistance, reduces corrosion of the metal coating and also loss of coating by electrolysis. Special precautions are necessary to exclude moisture from metallized-paper capacitors. See also 123 of 1950 (Wehe).

666.1.037.5 135

The Physical Aspect of Glass/Metal Sealing in the Electronic Valve Industry: Part 2—G. Trébuchon and J. Fieffer. (*Ann. Radioélect.*, vol. 5, pp. 243–258; July, 1950.) The processing technique for the glass alone is discussed: graphs show the optimum annealing temperature and duration, and optimum cooling rate, for glasses of different thicknesses and expansion coefficients. The effects of annealing on the quality of seal are discussed. Graphs based on polarimetric observations show the effects on the stresses produced in the seal in numerous cases. A table summarizes the effects of the different variables in the annealing cycle and also of the intrinsic properties of the materials. Part 1: 2253 of 1950.

669.15.26:666.1.037.5:621.385.832 136

Stainless Steel for Television—A. S. Rose. (*Metal Progress*, vol. 57, pp. 761–764; June, 1950.) The use of an alloy containing only 17 per cent Cr for the metal cones of large cathode-ray tubes for television results in a saving in cost compared with that for the alloy previously used, which contained 28 per cent Cr. The addition of small amounts of other metals is necessary to make the new alloy suitable for sealing to glass.

MATHEMATICS

512.31 137
On Certain Polynomials Introduced by Tchebycheff—H. Delange. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231, pp. 602–604; September 25, 1950.) A study of the asymptotic distribution of the zeros of P_n polynomials when n becomes infinitely great.

517.564.3:534.232 138
On the Extension of Some Lommel Integrals to Struve Functions with an Application to Acoustic Radiation—C. W. Horton. (*Jour. Math. Phys.*, vol. 29, pp. 31–37; April, 1950.)

517.564.3(083.5) 139
Tables of Integrals of Struve Functions—M. Abramowitz. (*Jour. Math. Phys.*, vol. 29, pp. 49–51; April, 1950.)

517.564.3(083.5) 140
A Short Table of Struve Functions and of Some Integrals Involving Bessel and Struve Functions—C. W. Horton. (*Jour. Math. Phys.*, vol. 29, pp. 56–58; April, 1950.)

519.272.15:621.392 141
Short Time Autocorrelation Functions and Power Spectra—R. M. Fano. (*Jour. Acoust. Soc. Amer.*, vol. 22, pp. 546–550; September, 1950.) The reciprocal relations between autocorrelation functions and power spectra, known as Wiener's theorem, are extended in a modified form to the case of experimental results obtained by means of filters with finite time-constants.

681.142 142
Comparison of Long-Time and Short-Time Analog Computers—V. Paschkis. (*Trans. AIEE*, vol. 68, Part I, pp. 70–73; 1949.)

681.142 143
Application of the California Institute of Technology Electric Analog Computer to Nonlinear Mechanics and Servomechanisms—G. D. McCann, C. H. Wilts, and B. N. Locanthi. (*Trans. AIEE*, vol. 68, Part I, pp. 652–660; 1949.) Full paper. Summary abstracted in 918 of 1950.

681.142:517.512.2 144
A New Fourier-Coefficient Harmonic Analyzer—S. Charp. (*Trans. AIEE*, vol. 68, PartI, pp. 644–649; Discussion 649–651; 1949.) Full paper. Summary abstracted in 674 of 1950.

681.142:621.392 145
A High-Speed Multiplier for Analogue Computers—Locanthi. (See 60.)

501:517.5:53 146
Formulas and Theorems for the Special Functions of Mathematical Physics [Book Review]—W. Magnus and F. Oberhettinger. Publishers: Chelsea Publishing Co., New York, 1949, 172 pp. (*Proc. Phys. Soc.*, vol. 63, p. 733 September, 1950.) This book, which is a reference work rather than a textbook, deals with Bessel functions, spherical harmonics, hypergeometric functions and elliptic functions, and, as special cases, Laguerre, Hermite and other functions. Chapters are included on integral transformations and on coordinate transformations.

517.43:[5+6 147
Operatorenrechnung und Laplacesche Transformation [Book Review]—K. W. Wagner Publishers: J. A. Barth, Leipzig, 1950, 471 pp., 42.80 DM. (*Tech. Mitt. schweiz. Telegr.-Teleph Verw.*, vol. 28, p. 334; August 1, 1950. In German.) Second revised edition of the book noted in 2827 of 1949.

517.9:[5+6 148
Die Differentialgleichungen der Technik und Physik (Differential Equations of Technology and Physics). [Book Review]—W. Hort and A. Thoma. Publishers: J. A. Barth, Leipzig 5th edn., 1950, 576 pp., 46.80 DM. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 335–336; August 1, 1950. In German.) This edition of "Die Differentialgleichungen des Ingenieurs," newly revised by Thoma, is a comprehensive textbook in eight parts dealing with the subject from elementary differential and integral calculus to differential and difference equations, Fourier series, variational calculus and integral equations. Graphical and mechanical methods of solution are described in addition to analytical methods.

"The book is clearly written and is recommended both for the beginner and the practising engineer."

517.91 149
Differential Equations [Book Review]—H. W. Reddick. Publishers: Wiley & Sons, New York, and Chapman & Hall, London, 2nd edn., 1949, 288 pp., 24s. (*Proc. Phys. Soc.*, vol. 63, p. 733; September 1, 1950.) A relatively elementary, but clear and accurate, textbook for intending engineers; partial differential equations are not dealt with.

MEASUREMENTS AND TEST GEAR

529.1:529.786 150
On a Periodic Fluctuation in the Length of the Day—H. F. Finch. (*Mon. Not. R. Astr. Soc.*, vol. 110, pp. 3–14; 1950.) From a study of the performance of a number of quartz-crystal clocks used in the Greenwich Time Service, and annual periodic fluctuation in the length of the day is deduced. The variation is of the order of ± 0.001 sec and has an accumulative effect in time of approximately ± 0.060 sec. This is in very close agreement with results obtained from independent data by N. Stoyko and demonstrates the persistent character of the phenomenon. See also 2275 of 1950 (Scheibe and Adelsberger).

621.3.011.5:621.365.55† 151
The Measurement of Dielectric Loss at High Frequencies and Under Changing Temperature—J. B. Whitehead and W. Rueggeberg. (*Trans. AIEE*, vol. 68, Part I, pp. 520–524; 1949.) Full paper. Summary abstracted in 137 of 1950.

621.317.088.4:621.314.12 152
The Fundamental Limitations of the Second-Harmonic Type of Magnetic Modulator as Applied to the Amplification of Small D.C. Signals—F. C. Williams and S. W. Noble. (*Proc. IEE* (London), Part II, vol. 97, pp. 445–459; Discussion, pp. 474–483; August, 1950.) The advantages of the second-harmonic type of magnetic modulator for the conversion of dc to ac are discussed and theoretical analysis is presented for an idealized modulator of this type, with particular reference to the influence of various controllable parameters on the signal-to-noise ratio and the zero error. Experimental work is described which provides qualitative verification of the theory when allowance is made for the assumption of a simplified B/H characteristic for the core material. Great care in the design of the various circuits is necessary to eliminate additional sources of noise and zero error. In apparatus described the noise output is mainly Barkhausen noise in the cores and is equivalent to a signal input of about 10^{-19} w for a bandwidth of 1 cps, the zero drift being equivalent to an input of about 3×10^{-18} w over a 2-hour period.

621.317.2:621.397.62 153
Television Laboratory Equipment—W. Werner. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 635–637; August 20, 1949. In English.) Paper presented at the International Television Conference, Zürich, 1948. Short descriptions of the special features and the uses of video signal generator, video distribution amplifier, high-frequency signal generator, microscope cro for observing any small portion of a television waveform, wide-band cro, sine-wave signal generator (up to at least 5 Mc), high-voltage voltmeter, film scanner, camera and studio-lighting equipment.

621.317.324(083.74)† 154
Two Standard Field-Strength Meters for Very-High Frequencies—D. D. King. (Proc. I.R.E., vol. 38, pp. 1048–1051; September, 1950.) "Methods of field-strength measurement are reviewed briefly and the design of field meters conforming closely to the conditions imposed by antenna theory is considered. Two instruments approaching ideal theoretical conditions and suitable for reference standards are described. The first of these contains an adjustable matching network. The second utilizes very fine wires on a styrofoam support."

621.317.353.3†:621.396.11:551.510.535 155
Ionospheric Cross-Modulation: Techniques of Measurement—C. C. Newton, F. J. Hyde, and H. G. Foster. (*Proc. Phys. Soc.*, vol. 63, pp. 616–623; August 1, 1950.) The techniques described were used for the investigations noted in 3219 of 1948 (Ratcliffe and Shaw), 194 of 1949 (Huxley, Foster, and Newton) and 1220 of 1950 (Huxley). The transferred modulation was deduced from receiver measurements of the carrier voltage and the af voltage of the wanted signal. The phase of the transferred modulation relative to that of the directly received disturbing signal was determined by forming a Lissajous figure; a phase changer was used to measure the phase difference. A development is described which permits modulation depth and phase to be displayed on a single cro.

621.317.411†+621.317.43 156
Measurement of Permeability and Magnetic Losses of Straight Samples—P. M. Prache and R. Cazenave. (*Câbles & Trans.* (Paris), vol. 4, pp. 216–233; July, 1950.) Advantages are gained by using a straight rod of the material under test in place of the toroidal sample normally introduced into the magnetic circuit. The calculation involved and the interpretation of results are simplified by considering the cylindrical sample replaced by an equivalent oblate spheroid and by using a coil much longer than this core so as to eliminate end effects. Practical inductance formulas are derived which may be used to determine permeability and loss coefficients.

621.317.443 157
A Permeameter for Magnetic Testing at Magnetizing Forces up to 300 Oersteds—R. L. Sanford and P. H. Winter. (*Bur. Stand. Jour. Res.*, vol. 45, pp. 17–21; July, 1950.) An instrument designed to test specimens up to 3 cm wide and 1 cm thick, with a preferred length of 28 cm. It is simpler and more rapid in operation than the Burrows permeameter and requires only a single specimen. Accuracy is within 1 per cent.

621.317.444† 158
Underwater Gaussmeter—L. Vérain and P. Jolivet. (*Rev. gén. Élect.*, vol. 59, pp. 405–408; September, 1950.) The instrument described comprises an air-driven rotor of special design located in the unknown field and having two collector brushes connected by line to a fluxmeter. The brushes are periodically short-circuited, so that the fluxmeter needle has a steady deflection proportional to the unknown field and independent of the duration of the deflection or the speed of rotation of the rotor. Results are reported of measurements made at Algiers, in 1940, of the vertical component of the earth's magnetic field at various depths in the vicinity of a destroyer.

621.317.723 159
A Simple Vibrating Condenser Electrometer—D. G. A. Thomas and H. W. Finch. (*Electronic Eng.*, vol. 22, pp. 395–399; September, 1950.) The dc input is converted to ac by applying it through a series resistor to the plates of a capacitor consisting of a stainless-steel reed vibrating at 550 cps close to a polished steel disk. The resultant alternating voltage is amplified, rectified and fed back to cancel the input voltage, the electrometer acting as a null detector. Diurnal zero drift is 1 mv on a full-scale sensitivity of 30 mv. An important application is to the measurement of ionization currents.

621.317.725 160
A New Expanded-Scale A.C. Voltmeter—N. P. Millar. (*Trans. AIEE*, vol. 68, Part I, pp. 641–643; 1949.) Full paper. Summary abstracted in 687 of 1950.

621.317.725.029.5 161
A Thermal Millivoltmeter for Measuring Radio-Frequency Voltages—N. Coulson. (*Proc. IEE* (London), Part III, vol. 97, pp. 344–348; September, 1950.) The measuring element is a thermocouple milliameter. The input impedance is varied by means of a turret switch which connects resistors in series or in parallel with the thermocouple, so that the output voltage of a source may be measured with various loads. At frequencies up to 100 Mc the error is less than 1 per cent for purely resistive 70-Ω sources, but may amount to 5 per cent when reactive elements are present.

621.317.727.025 162
The Polar Ammeter as an A.C. Potentiometer—The Synchropotentiometer—E. B. Brown. (*Jour. Sci. Instr.*, vol. 27, pp. 251–252; September, 1950.) Description of experiments showing that the voltage generated in the moving coil of a polar ammeter (1084 of 1948) can be varied both in phase and magnitude, so that the instrument can fulfil the functions of an ac potentiometer. An almost linear variation of amplitude can be obtained by moving the pointer over the scale. The phase can be varied by angular adjustment of the cradled synchronous motor. Results are quoted and a description is given of the apparatus.

621.317.73 163
A Direct-Reading Impedance-Measuring Instrument for the U.H.F. Range—W. R. Thurston. (*Gen. Radio. Exp.*, vol. 24, pp. 1–7; May, 1950.) This null-type instrument measures, on scales independent of frequency, conductances, and susceptances of either sign, from 1 to 400 millimhos at frequencies from 70 to 1,000 Mc. Three coaxial lines, one terminated by a resistance equal to its characteristic impedance, one a stub adjusted to $\lambda/8$ at the operating frequency and forming the susceptance standard, and the other connected to the unknown impedance, are fed at a common junction from a common source and have adjustable pickup loops which are so oriented that their combined output is zero. The loop-position scales are calibrated to read susceptance and conductance directly.

621.317.733.089.6:621.3.018.78† 164
A Method for Calibrating Distortion-Measurement Bridges—W. Hübner. (*Arch. tech. Messen*, pp. T73–T74; July, 1950.) Measurement of distortion with an ac bridge-comparator is liable to errors depending on the Q of the resonant circuit and especially on the order number of the harmonics present. In the calibration and test method described the measurement bridge is connected in the fourth arm of a resistance bridge to which, after balancing, a voltage of fundamental frequency is applied across one diagonal and a known harmonic voltage across the other diagonal. The distortion factor of the voltage appearing across the measurement bridge is thus known accurately.

621.317.755:621.3.015.3 165
Technique of Autosynchronous Observation of Transients—F. Lepri, I. F. Quercia, and B. Rispoli. (*Nuovo Cim.*, vol. 5, pp. 569–585; December 1, 1948.) Discussion of modifications to a circuit previously described *ibid.*, vol. 5, p. 384; 1948) to make the whole of the transient visible. The cro timebase is triggered by the transient, which passes through a delay line before being applied to the y-plates of the cro. Operation of the bootstrap, Miller and phantastron sawtooth-wave generators and the design of delay lines are considered.

621.317.755:621.317.791 166
Polar Vector Indicator—E. A. Walker, A. H. Waynick, and P. G. Sulzer. (*Trans. AIEE*, vol. 68, Part I, pp. 154–159; 1949.) See 2569 of 1949.

621.317.757 167
A Frequency-Spectrum Analyser for Radio Signals—J. Marique. (*HF, Brussels*, pp. 177–184; 1950.) An account is given of an instrument for analysis of signals from a distance. The principal features are described and other uses indicated. Various examples of measurements are illustrated.

621.317.78.029.64 168
The Measurement of Microwave Power at Wavelengths of 3 cm and 10 cm—R. Street and P. D. Whitaker. (*Proc. Phys. Soc.*, vol. 63, pp. 623–624; August 1, 1950.) A magnetron or klystron was connected via a waveguide to a matched-wedge constant-flow calorimeter. A directional coupler of known coupling factor was inserted in the waveguide. The absolute power delivered to a milliwattmeter matched to the low-power guide of the coupler can be calculated. For three types of instrument the ratios of absolute to indicated power were respectively 1.04, 1.04 and 1.10. The accuracy of the measurements is within about 2 per cent.

621.317.79:551.594.6 169
A Subjective Method of Measuring Radio Noise—H. A. Thomas. (*Proc. IEE*, (London), Part III, vol. 97, pp. 329–334; September, 1950.) The equipment described can be operated by inexperienced personnel. Sources of error are analyzed fully. Noise levels greater than 1 μv/m over the frequency range 2.5–20 Mc can be measured to within ± 5 db.

621.317.79:621.396.933 170
Monitoring Airways Radio—(*Wireless World*, vol. 56, p. 335; September, 1950.) A short account of the work carried out at the frequency-measurement station of the Ministry of Civil Aviation at Pailton, near Rugby, with illustrations of some of the equipment. Records are kept of all routine measurements and a monthly chart gives a day-to-day record of the frequencies of all the navigational beacons.

621.395.61.089.6 171
American Standard Method for the Pressure Calibration of Laboratory Standard Microphones: Z24.4—1949 (Abridged)—Beranek, Cook, Romanow, Wiener, and Bauer. (See 19.)

621.395.623.54.089.6 172
American Standard Method for the Coupler Calibration of Earphones: Z24.9—1949 (Abridged)—Beranek, Romanow, Morrical, Anderson, Bauer, Cook, and Wathen-Dunn. (See 21.)

OTHER APPLICATIONS OF RADIO AND ELECTRONS

534.321.9:061.3 173
Rome Ultrasonics Convention—Bradfield. (See 10.)

534.321.9.001.8:669.71:621.791 174
Ultrasonic Soldering of Aluminium—B. E. Noltingk and E. A. Neppiras. (*Nature*, (London), vol. 166, p. 615; October 7, 1950.) Experiments are described which show that in the process of tinning Al and its alloys by application of molten solder together with intense ultrasonic vibration, the action is that of removing the oxide skin by cavitation erosion. Relatively low frequencies were found preferable: a 50-w 18-kc oscillator was quite effective, while a 1-Mc oscillator supplied with 3 kv (rms) was ineffective.

538.569.2.047:621.38.001.8 175
Effects of Intense Microwave Radiation on Living Organisms—J. W. Clark. (Proc. I. R. E, vol. 38, pp. 1028–1032; September, 1950.) Radiation of wavelength about 10 cm was found the most dangerous. With much shorter waves, surface heating is produced and underlying tissues are little affected, while with much longer waves there is a general elevation of body temperature but no particular damage to the tissue. See also 2284 of 1949 (Salisbury, Clark, and Hines).

551.508.1:621.317.083.7 176
Automatic Range-Adjusting Radiosonde Recorder—G. E. Beggs, Jr. (*Trans. AIEE*, vol. 68, Part I, pp. 602–607; 1949.) Full paper. Summary noted in 703 of 1950.

621.317.755:531.771 177
An Electronic Tachometer—H. G. Jerrard and S. W. Punnett. (*Jour. Sci. Instr.*, vol. 27, pp. 244–245; September, 1950.) Details are given of an instrument capable of measuring the speed of revolution of a shaft, in any speed range, to within 0.05 per cent. The rotation of the shaft is made to generate an alternating voltage which, after amplification, is connected to the y-plates of a cathode-ray tube whose x-plates are connected to a variable-frequency oscillator.

621.365.54/.55† 178
The Design of H F. Generators for Industrial Use and the Development of their Application in France—J. Girardeau. (*Ann. Radioélect.*, vol. 5, pp. 259–275; July, 1950.) Illustrated account of many different applications of dielectric-loss and induction heating in industry, and of the equipment used.

621.384.611.2† 179
The Design of the Bevatron Magnet—D. Sewell. (*Elect. Eng.*, vol. 69, p. 721; August, 1950.) Summary of AIEE Summer General Meeting paper. The ring magnet for the Berkeley proton synchrotron will consist of four spaced quadrants, with an overall diameter of 135 ft. 9,700 tons of mild-steel plate will be required.

621.384.62† 180
Linear Accelerators—D. W. Fry and W. Walkinshaw. (*Rep. Progr. Phys.*, vol. 12, pp. 102–130; References, pp. 130–132;1948–49.) The axial and the radial stability of particles in radio-frequency fields are examined, basic types of particle accelerator are described and the traveling-wave and standing-wave types are considered in more detail. Methods of construction are described and also methods of applying radio-frequency power and of particle injection. Reported performance data of different linear electron accelerators are tabulated and possible future developments are outlined. Accelerators for heavy particles are discussed briefly.

621.385.833 181
Certain Properties of Electrostatic Fields encountered in Electron Lenses—P. A. Lindsay. (*Proc. Phys. Soc.*, vol. 63, pp. 699–702; September 1, 1950.) Fine details of the form of the equipotential lines in a bipotential electron

lens are revealed by application of the relaxation method to the solution of the Laplace equation in cylindrical co-ordinates. Asymmetry of the field between the adjacent ends of the two cylinders forming the lens is confirmed by the same method and is found to be of the order of 2 per cent in a particular case.

621.385.833 182
An Electron-Optical Apochromat—O. Scherzer. (*Z. Naturf.*, vol. 3a, pp. 544–545; 1948.) An example is calculated to show how chromatic aberration in electron lenses can be corrected by the use of plane metal foils permeable to electrons.

621.385.833 183
Reduction of the Spherical Aberration of Magnetic Electron Lenses—U. F. Gianola. (*Proc. Phys. Soc.*, vol. 63, pp. 703–708; September 1, 1950.) Discussion of a method for increasing the resolving power of asymmetrical magnetic electron lenses; the lens field is reinforced with the field of a small air-cored coil.

621.385.833 184
Second-Order Beam Focusing of Charged Particles in Homogeneous Magnetic Fields—H. Hintenberger. (*Z. Naturf.*, vol. 3a, pp. 669–670; 1948.

621.385.833 185
A Removable Intermediate Lens for Extending the Magnification Range of an Electron Microscope—J. Hillier. (*Jour. Appl. Phys.* vol. 21, pp. 785–790; August, 1950.) A lens is described which increases the ratio of maximum to minimum magnification of a conventional instrument to 25 to 1 without affecting the accessibility of the objective and projection lens pole pieces, thus making available the low magnifications needed for survey purposes, and so forth.

621.387.4† 186
Hydrogen-Filled Geiger Counters—B. Collinge. (*Proc. Phys. Soc.*, vol. 63, pp. 665–674; September 1, 1950.) A new quench circuit is described and the characteristics of counters with permanent-gas filling are discussed in detail.

621.387.4† 187
Temperature Dependence of Counter Characteristics in Self-Quenching Geiger-Müller Counters—W. R. Loosemore and D. Taylor. (*Proc. Phys. Soc.*, vol. 63, pp. 728–729; September 1, 1950.) Comment on 1983 of 1950 (Parkash and Kapur).

621.387.4† 188
After-Effects in Ultraviolet-Sensitive Counters—H. Neuert. (*Z. Naturf.*, vol. 3a, pp. 221–225; 1948.)

681.142:533.6 189
Electric Analogue Computing Techniques for Complex Vibration and Aeroelestic Problems—G. D. McCann and R. H. MacNeal. (*Elec. Eng.*, vol. 69, p. 724; August, 1950.) Summary of AIEE Summer General Meeting paper.

621.384.6† 190
The Acceleration of Particles to High Energies [Book Review]—Publishers: Institute of Physics, London, 58 pp., 10s.6d. (*Jour. Sci. Instr.*, vol. 27, p. 255; September, 1950.) This volume in the Institute's 'Physics in Industry' series is based on papers presented at the Institute's 1949 Convention at Buxton.

621.385.833 191
The Practice of Electron Microscopy [Book Review]—D. G. Drummond (Ed.). Publishers: Royal Microscopical Society, London, pp. 141, 21s. (*Jour. Sci. Instr.*, vol. 27, p. 255; September, 1950.) A comprehensive treatise on the detailed techniques used. It was produced by a group of twelve members of the Electron Microscopy Group of the Institute of Physics.

PROPAGATION OF WAVES

538.566 192
Formulation of Huygens' Principle—W. Franz. (*Z. Naturf.*, vol. 3a, pp. 500–506; 1948.) Making use of Green's dyad, a formulation of Huyghens' principle for em waves is derived which, like Kirchhoff's scalar formula, makes it clear that for selected boundary values the wave equations are satisfied. Kirchhoff's theory does not solve a boundary-value problem but a discontinuity problem. In contradistinction to Kottler, the discontinuity is regarded as basic to Kirchhoff's theory and not as a property of the 'black' screen.

538.566:551.510.535 193
The Poynting Vector in the Ionosphere—J. C. W. Scott. (Proc. I.R.E., vol. 38, pp. 1057–1068; September, 1950.) Formulas and curves are given for calculating the polarization and complex Poynting vector of a radio wave in the ionosphere in the most general case. Deductions are made concerning the direction of energy flow for the ordinary and extraordinary modes in a parabolic distribution of ionization, for vertical incidence. When collision is taken into account the deflection from the vertical has a small westward component for both modes. The normal ionization gradient with latitude, together with diurnal changes in the ionized region, can explain the diurnal variation in the $f_x - f_0$ critical-frequency difference as due to the variation in the total path deflection. See also 2516 of 1949.

538.566:621.396.67 194
Cylindrically Diverging Electromagnetic Waves in a Medium with Nonuniform Electrical Properties (Elias-Layer) above a Semiconducting Earth—van der Wyck. (See 32.)

621.396.11:523.74/.75 195
Solar Notes—Newton. (See 106.)

621.396.11:621.317.353.3† 196
Gyro-interaction—Please note that the above UDC number will be used in future instead of 621.396.812.

621.396.11:621.317.353.3† 197
Variability of the Resonance Frequency in Gyro-interaction of Radio Waves—M. Carlevaro. (*Nuovo Cim.*, vol. 5, pp. 535–550; December 1, 1948.) It is suggested that small variations of ionic density in the E layer satisfactorily account for the variations in resonance frequency found experimentally (513 of 1947 and 2328 of 1950). Theoretical calculations predict a smaller range of variation than that observed, unless it is assumed that in the lower levels of the E layer the electronic density is reduced mainly by negative-ion formation, while in the upper levels it is reduced by recombination with positive ions.

621.396.11:621.317.353.3†:551.510.535 198
Ionospheric Cross-Modulation: Techniques of Measurement—Newton, Hyde, and Foster. (See 155.)

621.396.11.029.62 199
Experimental Study of the Propagation of Metre Waves: Measurements in Aircraft—H. Vigneron. (*HF, Brussels*, pp. 191–198; 1950.) Conflicting theories of propagation are discussed. The field calculated according to the ray theory has a series of maxima and minima, and is limited to the optical range. Theories based on diffraction give a field decreasing progressively with increasing distance and extending beyond the optical range. Measurements were made of 116.1-Mc signals received in aircraft at heights of 300, 1,500 and 3,000 m as the transmitter was approached. At great altitudes results conform to the ray theory; at small altitudes and well below the line of sight, van der Pol curves are applicable, with a gain with height which can only be determined by systematic experiments.

621.396.81.029.63/.64:551.5 200
A Radio Meteorological Investigation in the South Island of New Zealand—B. Milnes and R. S. Unwin. (*Proc. Phys. Soc.*, vol. 63, pp. 595–616; August 1, 1950.) Local conditions in New Zealand, particularly with off-shore föhn winds, are favourable to the formation of radio ducts. Modern techniques were used to explore thoroughly the atmospheric conditions over a range of 200 km from the coast and up to 600 m above sea level. The results, obtained over the period from September 1946 to the end of 1947, are displayed as isopleths of the atmospheric parameters and of refractive index. Contours of radio field-strength are shown for selected days for wavelengths of 300 cm, 60 cm, 10 cm, and 3 cm. The properties of the ducts and the experimental technique are fully described.

621.396.812.3:551.510.535 201
The Fading of Radio Waves of Medium and High Frequencies—R. W. E. McNicol. (*Proc. IEE* (London), Part III, vol. 97, p. 366; September, 1950.) Discussion on 443 of 1950.

621.396.812.4.029.64 202
Microwave Propagation Experiments—L. E. Thompson. (*Proc. I.R.E.*, (Australia), vol. 11, pp. 204–209; August, 1950.) Reprint. See 2894 of 1948.

RECEPTION

551.594.6 203
Some Measurements of Atmospheric Noise at High Frequencies—H. A. Thomas. (*Proc. IEE*, (London), Part III, vol. 97, pp. 335–343; September, 1950.) Measurements of atmospheric noise at a number of stations in various parts of the world have been made since 1922 over periods ranging from a few days to several years. A summary of these observations has previously been given [534 of 1948 (Thomas & Burgess)]. The frequencies used were 2.5, 5, 10, 15 and 20 Mc, and the method was based on aural comparison of the received noise and a locally generated noise signal of controllable amplitude (see 169 above). Some of the results obtained are presented in the form of mean values for each month and for each hour of the day, together with figures indicating the degree of scatter about the mean value. The characteristics of atmospheric noise at one location do not appear to be applicable over a large area; this and other considerations throw doubt on the concept that lightning discharges are practically the sole source of high-frequency noise. It is suggested tentatively that some noise sources may be quite local, but more experimental data are required to confirm or disprove this.

621.396.621:621.396.619.11 204
The Synchrodyne as a Precision Demodulator—D. G. Tucker and R. A. Seymour. (*Wireless Eng.*, vol. 27, pp. 227–237; August and September, 1950.) The synchrodyne circuit, in which the modulation frequency is extracted by filtration after the detector, may lead to distortion due to phase modulation of the local oscillator at the modulation frequency. An analysis of this distortion in the basic circuit is given, and practical methods of achieving precision, stability and freedom from distortion are described. Phase modulation is limited by the use of tube reactors, and constancy of gain is achieved by application of negative feedback.

621.396.823:537.523.3 205
Radio Influence from High-Voltage Corona—G. R. Slemon. (*Trans. AIEE*, vol. 68, Part I, pp. 198–204; Discussion, pp. 204–205; 1949.) See 1780 of 1949.

621.396.828 206
Reduction of Interference from Radio-Frequency Heating Equipment—G. W. Klingaman. (*Trans. AIEE*, vol. 68, Part I, pp. 718–724; Discussion, p. 724; 1949.) Discussion of the causes of the generation of very high frequencies by radio-frequency heating equipment, and of measures for its reduction, particular attention being given to harmonic suppression and effective screening.

621.396.933:621.395.625.2 207
An Automatic Monitoring Recorder—(*Engineer*, (London), vol. 190, p. 186; August, 18, 1950.) Short description of equipment for continuous recording of speech signals transmitted from an aircraft to a ground control station. The speech-frequency range is limited to 500–3,000 cps. Recording is on standard Kodak film, a sapphire needle producing lateral indents, without cutting. The recording head is traversed across the film to obtain 120 sound tracks on each side, so that 120 ft of film suffice for a 24-hour period. Any portion of the record can be reproduced without interrupting the recording.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 208
Photons and Waves—D. Gabor. (*Nature*, (London), vol. 166, pp. 724–727; October 28, 1950.) Abstract of parts of a lecture delivered in Paris, May 9, 1950. on 'La Théorie des Communications et la Physique.' A comparison is made of the quantum and classical methods of describing signals. The 'information cell' is taken as a convenient unit for discussing communication problems; by the classical method this has two data associated with it, an amplitude and a phase, but by the quantum method only one datum, of the nature of an amplitude. More information, however, is gained by the latter method since the total number of distinguishable steps of the single datum is greater than the product of the numbers of distinguishable steps of the two data in the classical analysis. The theory is illustrated by a determination of the optimum conditions for interchange of energy between a weak signal and a transverse electron beam in a waveguide.

621.39.001.11:535.42 209
Diffraction and Quantity of Information—A. Blanc-Lapierre and M. Perrot. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 231, pp. 539–541; September 11, 1950.) The system considered is that constituted by an aperture, an object at infinity composed of incoherent sources, and a diffraction image at infinity. From the correspondence between image and object, the quantity of information transmitted by the aperture is deduced.

621.39.001.11:621.317.35 210
Signals with Limited Spectra and their Transformations—J. Oswald. (*Câbles & Trans.* (Paris), vol. 4, pp. 197–215; July, 1950.) Detailed mathematical treatment. See also 3069 of 1949.

621.395.44:621.315.052.63 211
Line Tuning Equipment Used with Coaxial Cable for Carrier-Current Installation on Power Lines—H. J. Sutton. (*Trans. AIEE*, vol. 68, Part I, pp. 44–48; Discussion, pp. 48–49; 1949.)

621.396.61/.62 212
A Frequency-Modulated Transmitter-Receiver for Motor Cycles—(*Engineer* (London), vol. 190, p. 172; August 18, 1950.) A 27-tube equipment in two units mounted on either side of the rear wheel. It is crystal controlled and has an FM radio-frequency output of 10 w on a spot frequency in the band 68–100 Mc. Sensitivity is 1 μv carrier input for 10-db quieting. 'Standby' power consumption is only 18 w. The equipment can be used within the temperature range −40°C to 70°C. In conjunction with a 20-w control transmitter it has a specified service radius of 20 miles. A selective calling system enables any one, or all, of 90 such units to be called from the control station.

621.396.619:621.392.52 213
Polyphase Modulation as a Solution of Certain Filtration Problems in Telecommunication—I. F. Macdiarmid and D. G. Tucker. (*Proc. IEE*, Part III, vol. 97, pp. 349–358; September, 1950.) An important class of filtration problems in telecommunication is associated with frequency changing; it includes the generation and demodulation of single-sideband carrier channels and the elimination of image-frequency interference in heterodyne demodulators, such as the superheterodyne radio receiver or the conventional wave analyzer. Filters for these applications are often difficult to design or realize, or may be inconvenient on account of variable tuning, and so forth.

Polyphase modulation can be used as part of the frequency-changing process with great advantage. It can eliminate the need for difficult or inconvenient filters, although other design problems are introduced which may sometimes be as difficult to solve. The basis of the advantages given by polyphase working is that polyphase signals possess an identifying property additional to that of frequency, namely sequence. By using circuits which distinguish between signals of the same frequency but opposite sequence, it is possible, without any preliminary filtration, to separate signals which lie in the same frequency band after modulation, and which, therefore, could be separated by normal means only by filters before the modulation stage.

The first section of the paper outlines the main filtration problems which can be tackled by polyphase methods, and then the necessary polyphase theory is given. This is followed by a discussion of circuit design for polyphase modulation and sequence discrimination. The list of 30 references shows that there have been many publications covering some of the separate applications of this work, but the paper is believed to present for the first time a comprehensive theory of polyphase modulation embracing all the known applications.

621.396.619.16:621.396.41 214
A Time Division Multiplexing System—W. P. Boothroyd and E. M. Creamer, Jr. (*Trans. AIEE*, vol. 68, Part I, pp. 92–97; 1949.) See 3258 of 1949.

621.396.65 215
A Microwave Communication Relay System—W. P. Boothroyd and H. J. Churchill. (*Trans. AIEE*, vol. 68, Part I, pp. 637–641; 1949.) An incoming FM radio-frequency carrier causes deviation of the output of a local oscillator in accordance with the modulation of the incoming carrier, the oscillator output being amplified and radiated as the repeated signal. The whole system constitutes a negative-feedback amplifier and a study of its performance as a repeater is made on this basis. The extreme simplicity of its electrical and mechanical design makes the use of a tower unnecessary, a simple pole being sufficient to support the repeater equipment.

621.396.65 216
Problems to be Solved in the Application of Microwave Equipment—R. C. Cheek. (*Elec. Eng.*, vol. 69, p. 718; August, 1950.) Summary of AIEE Summer General Meeting paper. Points to be dealt with in establishing a microwave channel include determination of frequency band, selection of terminal sites, and calculation of inherent losses.

621.396.65:621.311 217
Microwave Channels for Power System Applications—(*Trans. AIEE*, vol. 68, Part I, pp. 40–42; discussion pp. 42–43; 1949.) An AIEE committee report discussing the advantages and disadvantages of microwave links for telemetry, supervisory control, and communication.

621.396.65:621.396.43.029.6 218
Radio-Beam System Planning—W. Gerber. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 648–650; August 20, 1949. In German.) Paper presented at the International Television Conference, Zürich, 1948. Discussion of the possible international extension of the present Swiss system of high-altitude stations.

621.396.65.029.6.001.4(494) 219
Directional Transmission Tests in the Alps Contributing to the Establishment of a Swiss R/T Network—W. K. Pein. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 303–317; August 1, 1950.) French and Italian versions based on paper abstracted in 2326 of 1949.

621.396.65.029.64 220
Passive Relay Stations of the Afourer/Bin-el-Ouidane Link—R. Chaux and J. Dascotte. (*Ann. Radioélec.*, vol. 5, pp. 220–229; July, 1950.) Expressions for the radiation pattern and gain of a plane metallic mirror are derived. The link considered is between two stations in Morocco 15 km apart in mountainous country. Two intermediate relay stations are used. At one a plane duralumin mirror of area 10 m² is mounted in a rigid frame 2 m above the ground. At the other, two similar plane mirrors are supported in an open cage on a 40-m pylon. Both mirrors are hinged to facilitate adjustment. Terminal transmitter and receiver antennas use parabolic mirrors of 10-m² aperture. Polarization is vertical. $\lambda=9.5$ cm. Using a 1-w FM signal and a 1.5-Mc passband in the receivers, communication has been maintained since 1949 in extreme climatic conditions. Calculated field strength and noise level are in fair agreement with actual values.

621.396.932 221
Liverpool Harbour Communications—(*Wireless World*, vol. 56, pp. 277–279; August, 1950.) See also 2633 of 1950.

SUBSIDIARY APPARATUS

621–526 222
Comparison of Steady-State and Transient Performance of Servomechanisms—H. Chestnut and R. W. Mayer. (*Trans. AIEE*, vol. 68, Part I, pp. 765–777; 1949.)

621–526 223
Instrument Inaccuracies in Feed-Back Control Systems with Particular Reference to Backlash—H. T. March, M. Yachter, and J. Zauderer. (*Trans. AIEE*, vol. 68, Part I, pp. 778–788; 1949.)

621–526 224
Analyzing Contactor Servomechanisms by Frequency-Response Methods—J. Kochenburger. (*Elec. Eng.*, vol. 69, pp. 697–692; August, 1950.) An approximation method which facilitates the selection of compensating networks for improving the performance of contactor servomechanisms.

621.314.6 225
New Developments in Rectifier Technique—F. Kesselring. (*Tech. Mitt. schweiz. Telegr.-

Teleph. Verw., vol. 28, pp. 297–303; August 1, 1950. In French and German.) Description of two types of rectifier. The first is a vibrator mechanism. The interrupter tongues are prism-shaped and weigh about 60 mg; enclosed in a sealed container with inert gas under pressure they can withstand over 10 kv. Models developed include a 200-A and 1,000-A type. The second design is a grid-controlled rectifier tube with Cs-vapor filling. Operating voltage is low; peak voltage 800–3,000 v. Tubes passing 30 A have been constructed; a 150-A type is under development.

621.316.578.1 226
Electrical Timing Devices—F. E. Reeves. (*Elec. Mfg.*, vol. 42, pp. 114–119, 168; September, 1948.) Ten points to be considered in the selection and application of electrical switch timers are enumerated and a chart is given outlining the operating characteristics of commercially available equipment.

621.316.722.1 227
A New Precision A.C. Voltage Stabilizer—G. N. Patchett. (*Proc. IEE*, Part II, vol. 97, No. 58, pp. 529–538; discussion, pp. 538–540; August, 1950.) Various types of stabilizer are discussed and an account is given of the design and performance of a stabilizer for meter testing applications which uses a temperature-compensated thermistor bridge. The stabilization ratio for a 10-v change of mains voltage (230 v nominal) is about 1,100. An output up to 2 kva can be obtained.

621.316.722.1 228
A Simple Form of Voltage Stabilizer—N. K. Saha, B. S. Chandersekhara, and M. K. Sundaresen. (*Proc. Nat. Inst. Sci.* (India), vol. 16, pp. 127–133; March and April, 1950.) Operating in the range 600–2,000 v and suitable for Geiger-Müller counters, the stabilizer consists of an air-discharge tube, under variable pressure and connected, in series with a 30–40-MΩ resistor, in parallel with the rectified output of a transformer. The voltage drop across the tube remains constant, for a given pressure, over a wide range of transformer output voltage, owing to gas ionization, the stabilized voltage increasing with pressure. A deviation of about 2.5 per cent at a stabilized voltage of 1,500 v is quoted for transformer-output variation from 2,000 to 3,500 v.

621.316.722.1 229
Voltage Stabilizers assure Top Performance—(*Elec. Mfg.*, vol. 42, pp. 108–113 . . . 192; September, 1948.) Operating principles and performance data for magnetic, electronic, and servomechanism types of equipment for supplying constant-voltage power.

621.316.722.1 230
A Modified Moving-Coil Voltage Regulator of High Sensitivity—N. W. W. Ellis. (*Jour. Sci. Instr.*, vol. 27, pp. 248–249; September, 1950.) "Modifications to a standard commercial voltage regulator are described, which result in stabilization of a mains supply line to within ± 0.1 v for changes of load or input supply voltage and frequency within the ranges normally encountered. Loads of up to 7.5 kva may be applied."

621.316.722.1.076.7 231
The Cathode Follower as a Voltage Regulator—A. P. Willmore. (*Electronic Eng.* (London), vol. 22, pp. 399–400; September, 1950.) Since the output voltage developed across the cathode load resistor is proportional to $V_g + V_a/\mu$, where μ is the amplification factor, fluctuations of V_a are reduced by a factor depending on μ. The cathode follower may be used (a) to increase the current range over which stabilization is satisfactory with a particular voltage-reference tube, (b) to provide a high reference potential for a series-parallel type of voltage regulator.

621.352/.355 232
Special Purpose Batteries—A. Fischbach. (*Elec. Eng.*, vol. 69, pp. 701–704; August, 1950.) Three types developed for service use are discussed.

621.355 233
Electric Batteries: Recent Patents—L. Jumau. (*Rev. gén. Elect.*, vol. 59, pp. 372–378; September, 1950.) Developments in primary batteries, chiefly of the dry type, are reviewed. See also 3177 of 1950.

621.396.683:621.396.65 234
Power Supplies for Microwave Relay Systems—H. M. Ward. (*Trans. AIEE*, vol. 68, Part I, pp. 631–636; 1949.) Interruption of the main ac supply causes the load to be transferred to a battery-operated vibrator in under 0.1 second. This is cut out when a petrol-electric set, whose start is delayed 15 seconds to avoid unnecessary starting during very short power failures, reaches a steady operating condition. The various units of the equipment are described. Performance data for various radio-beam links indicate the high degree of reliability achieved.

771.36:537.228.4 235
An Electro-optical Shutter for Photographic Purposes—A. M. Zarem, F. R. Marshall, and F. L. Poole. (*Trans. AIEE*, vol. 68, Part I, pp. 84–91; 1949.) A description of the development of a simple and reliable optical shutter using a Kerr cell as a light tube, with which photographic studies of electric discharges have been made using an effective exposure time of 0.04 μ second. The control can be made sufficiently positive and accurate to permit initiation of operation at any preselected instant to within about 0.005 μ second.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2 236
Considerations on Facsimile Transmission Speed—H. F. Burkhard. (*Trans. AIEE*, vol. 68, Part I, pp. 418–423; discussion, p. 423; 1949.) A résumé of the work of many investigators and mathematical analysis of the factors which limit the speed of facsimile transmission in various systems. A method capable of transmitting 640 square inches of copy per minute over a channel 192 kc wide, or 160 square inches with a channel width of 48 kc, is described.

621.397.2 237
New Facsimile System—M. Frank. (*Ann. Geofis.*, vol. 2, pp. 532–544; October, 1949.) Suitable for transmission of weather maps, graphs, and printed matter of size up to 25 cm×30 cm, by telephone line or radio link. An electromechanical recording system is used, the modulated subcarrier being obtained by interrupting the scanning beam at 3 kc. The number of scanning lines can be 4 per mm or less. The subcarrier frequency, after frequency division, is used to synchronize the movements of transmitter and receiver drums. Several copies can be produced simultaneously at the receiving end.

621.397.24/.26 238
Long-Distance Television Links between Fixed Points—F. Vecchiacchi. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 647–648; August 20, 1949. In French.) Paper presented at the International Television Conference, Zürich, 1948. A short discussion of the economics of cable and radio links, with examples of both types at present in use.

621.397.24 239
Television Distribution over Short Wire Lines—P. Adorian. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 650–653; August 20, 1949. In English.) Paper presented at the International Television Conference, Zürich, 1948. See 2339 of 1949.

621.397.24.018.78†:621.315.212 240
Characteristics of Coaxial Pairs at Frequencies Involved in High-Definition Television Transmission—Fuchs. (See 27.)

621.397.26:621.396.615.142.2:621.396.621.53 241
The Klystron Mixer Applied to Television Relaying—Learned. (See 275.)

621.397.26:629.135 242
First Results of Stratovision Tests in the United States of America—E. J. Aubort. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 653–657; August 20, 1949. In French.) Paper presented at the International Television Conference, Zürich, 1948. An account of tests carried out near Pittsburg with the relay aircraft at a height of 8,000 m, when the useful ground range exceeded 400 km. A map shows corresponding ranges in Europe for an aircraft at the same height over Zürich. A second map indicates the possibilities of international program exchange in Europe, using seven aircraft and taking account of the coaxial cable envisaged by the CCIF for 1952. See also 3801 of 1946, 3279 of 1947 (Nobles) and 233 of 1949 (Sleeper).

621.397.331.2 243
'Knight' Scanning, Method giving Improvement of Television Picture Definition without Increase of Bandwidth—P. M. G. Toulon. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 638–641; August 20, 1949. In French.) Paper presented at the International Television Conference, Zurich, 1948. Summary abstracted in 870 of 1949.

621.397.331.2:778.5 244
Notes on [Picture] Analysis in Television with Continuously Moving Film—S. Mallein. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 603–605; August 20, 1949. In French.) Paper presented at the International Television Conference, Zürich, 1948.

621.397.331.2:778.5 245
Luminescent-Screen 875-Line Scanning of Film Pictures—A. Karolus. (*Bull schweiz. elektrotech. Ver.*, vol. 40, pp. 566–569; August 20, 1949, In German.) Paper presented at the International Television Conference, Zürich, 1948. A description of the apparatus, with a diagram illustrating the general layout, is given. The cathode-ray tube used is identical in construction with the usual type of projection tube and is operated at 25–40 kv. The raster surface is 6 cm×9 cm and the screen is of the low-persistence type. Experiments showed that a ZnO phosphor gave a much higher degree of modulation than other phosphors tested. Operation of the equipment, using a flat type of photocell with a semitransparent photo layer about 6 cm in diameter, was satisfactory for both carrier-frequency and low-frequency scanning.

621.397.335 246
New Possibilities for External Synchronization of Home-Television Pictures—W. Gerber. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 646–647; August 20, 1949. In German.) Paper presented at the International Television Conference, Zürich, 1948.

621.397.5 247
Work towards International Television—R. Barthelemy. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 535–538; August 20, 1949. In French.) Paper presented at the International Television Conference, Zürich, 1948. Discussion of the various problems involved, par-

ticularly the choice of the line standard and the question of interlacing.

621.397.5 248
Evolution of Television—A. Ory. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 545–547; August 20, 1949. In French.) Paper presented at the International Television Conference, Zurich, 1948.

621.397.5 249
The Present Status of Color Television—(Proc. I.R.E., vol. 38, pp. 980–1002; September, 1950.) The report of the Senate Advisory Committee on Color Television, given in full. A bandwidth of 6 Mc is considered adequate, representing an optimum compromise between quality and quantity of service. The general principles of the CTI line-sequential system, the CBS field-sequential systems involving line and dot interlace, and the RCA system employing a dot-sequential system with the method of mixed highs to increase definition, are explained in detail and performance characteristics, such as color fidelity, flicker, resolution and break-up of the picture for moving objects, are tabulated for each system. Appendices reproduce official correspondence concerning the report, and also the results of tests of flicker and color fidelity by the National Bureau of Standards.

621.397.5 250
Mixed Highs in Color Television—A. V. Bedford. (Proc. I.R.E., vol. 38, pp. 1003–1009; September, 1950.) Tests on the human eye, using projected color-test slides, are described in detail; these indicate that the acuity for resolving color differences is less than half as great as that for differences in brightness, so that the bandwidth used for the color transmissions can be correspondingly reduced. In the system described, of three color signals each use a 2-Mc frequency band and the red and green 'mixed highs,' which represent the brightness values, use an additional 2-Mc band, so that the total video bandwidth required will be 8 Mc, compared with 12 Mc that would be required if all the three colors required bandwidths of 4 Mc. The dot-interlace method, as used in the latest RCA system, reduces the required bandwidth to 6 Mc.

621.397.5:534.321.9 251
Underwater Television by means of Ultrasonics—M. Federici. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 657–659; August 20, 1949. In Italian.) Paper presented at the International Television Conference, Zürich, 1948. Discussion of a possible system which should be quite practical for short-distance transmissions.

621.397.5:535.317.9 252
The Schmidt Optical System—Rinia. (See 82.)

621.397.5:621.38 253
Electronics in Television—V. K. Zworykin. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 549–560; August 20, 1949. In English.) Paper presented at the International Television Conference, Zürich, 1948. General discussion of the subject, with special reference to recent developments in electronic camera tubes, direct-viewing receiver tubes, and color television.

621.397.5:778.5 254
Recording of Television Transmissions on Film with a view to International Exchange of Programmes—Y. L. Delbord. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, pp. 571–580; August 20, 1949. In French.) Paper presented at the International Television Conference, Zürich, 1948. A detailed discussion of the various problems involved, description of different methods and types of equipment, and illustrations of the quality of picture possible with such systems. Summary noted in 877 of 1949.

621.397.62:621.317.2 255
Television Laboratory Equipment—Werner. (See 153.)

TRANSMISSION

621.396.615.142.2:621.396.621.53:621.397.26 256
The Klystron Mixer Applied to Television Relaying—V. Learned. (Proc. I.R.E., vol. 38, No. 9, pp. 1033–1035; September, 1950.) The phase-modulation sidebands obtained from a klystron amplifier with beam-voltage modulation are used to give a mixing action, and the output resonator is used to select one of the sidebands. The construction and operating details are given for a klystron amplifier, Type SAC-19, with a mixer output of 1 w over a 20-Mc band centered at about 6,000 Mc.

TUBES AND THERMIONICS

533.723+621.396.822 257
Spontaneous Fluctuations—MacDonald. (See 79.)

621.314.65/.67 258
Fundamental Processes in Gaseous Tube Rectifiers—A. W. Hull. (*Elec. Eng.*, vol. 69, pp. 695–700; August, 1950.) Phenomena in gas-filled tubes are considered in relation to the laws of thermionic emission and electron space charge.

621.383.4 259
Photoconductive Cells of Cadmium Selenide—E. Schwarz. (*Proc. Phys. Soc.* (London), vol. 63, pp. 624–625; August 1, 1950.) Operating characteristics are given for a cell produced by methods described in 1102 of 1949. The theory given in 3578 of 1949 is supported by the present experimental results.

621.385.029.63/.65 260
The Anticyclotron, a New Type of Traveling-Wave Valve with Magnetic Field—G. Mourier. (*Ann. Radioélec.*, vol. 5, pp. 206–219; July, 1950.) Steady-state conditions possible in tubes with magnetic fields are discussed. The principle of the projected tube, which is ring shaped and has no radial field, is analogous to that of the cyclotron; but in this case the electron beam is slowed down as a result of its synchronism with a retarded traveling wave. The interaction of the two may be compared to the oscillations in a magnetron. A first approximation based on Doehler's theory (250, 261 and 1544 of 1949) shows the theoretical gain to be of the same order as in a helix traveling-wave tube, while the efficiency and dc input may be much higher. The induction field required is comparatively very small. According to the relative value of this field, the tube behaves like a linear traveling-wave tube of the Kompfner-Pierce type or like one with a transverse magnetic field. On account of the low field-strength required, the anticyclotron may possibly be used to generate mm waves.

621.385.029.63/.64 261
Small-Signal Theory of Wave Propagation in a Uniform Electron Beam—G. G. Macfarlane and A. M. Woodward. (*Proc. IEE*, Part III, vol. 97, pp. 322–329; September, 1950.) Analysis is presented for the three systems constituted by a planar uniform beam (a) between conducting sheets, (b) in free space, (c) between reactive-impedance sheets, the cast being a simple form of traveling-wave tube. For small signals, the em field in a traveling-wave tube may be split up into an infinite set of modes and for each mode there can be two forward and two reverse waves. The amount of each mode present depends on the method of excitation. At high frequencies the maximum amplification occurs when the ratio of beam velocity to phase velocity is slightly greater than unity.

621.385.029.63/.64:537.525.92 262
On Certain Effects of the Space Charge in Traveling-Wave Valves—R. Berterottière and G. Convert. (*Ann. Radioélec.*, vol. 5, pp. 168–178; July, 1950.) The effects of space charge are investigated theoretically by introducing into the equations for the electron dispersion a complex coefficient analogous to the coupling resistance function for the beam and the field. Electron trajectories are assumed to be rectilinear. For signals of small amplitude the theory may be applied up to the limiting case of a very weak focusing field, when space-charge effects are negligible. In the case of signals of large amplitude the effects are complex; they may result in increased efficiency.

621.385.032.216 263
The Barium-Oxide-on-Tungsten Cathode Interface—E. B. Hensley and J. H. Affleck. (*Jour. Appl. Phys.*, vol. 21, pp. 938–939; September, 1950.) The compound formed at the interface between a w cathode base and its BaO coating has been identified, by X-ray diffraction, as principally $BawO_3$. Corresponding tungstites are found when SrO or the solid solution (BaSr)O is used for the coating.

621.385.032.24:537.311.315 264
Variations of Grid Contact Potential and Associated Grid Currents—H. B. Michaelson. (*Jour. Frank. Inst.*, vol. 249, pp. 455–473; June, 1950.) A review of the subject, with a comprehensive bibliography. The grid-cathode Volta potential, or 'true contact potential,' is shown to be the difference between the work functions of the grid and cathode; it differs essentially from the quantity called 'contact potential' that is generally measured in routine tests of tubes. When the potential of the grid is negative, the current in the external grid circuit consists of several small currents due to various causes which change during the life of the tube and thus alter the tube characteristics. Methods that have been suggested for controlling these changes are outlined. Thermionic work functions are listed for 44 pure metals, for monolayers of various substances on Ni, Mo or w bases, and also for oxide coatings. Another list gives the effect of various gases and vapors on the work function of 15 metals.

621.385.4 265
The Internal Resistance of a Pentode—J. L. H. Jonker. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 179–194; July and September, 1950.) For output pentodes the main factor determining the internal resistance is the direct effect of anode voltage on cathode current. For high-frequency pentodes two further effects are of importance, viz., (*a*) the absorption by the screen grid of electrons repelled by the suppressor, and (*b*) the absorption by the screen grid of electrons reflected by the anode and transmitted by the suppressor. Both these effects again depend on the anode voltage. Measured values of the resistance are compared with values calculated from theory, and the discrepancies are related to the simplifying assumptions made regarding the operation of the tube.

MISCELLANEOUS

621.396:061.4 266
The 17th National Radio Exhibition, Castle Bromwich, September 6 to 16, 1950—(*Electronic Eng.* (London), vol. 22, pp. 378–384; September, 1950.) Brief descriptions of selected radio and television equipment and accessories, compiled from information supplied by the manufacturers.

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HI-Q CERAMIC DISK CAPACITORS

HI-Q Ceramic Disk Capacitors for by-passing, blocking, or coupling are being used by the millions by television receiver manufacturers who demand the utmost in performance.

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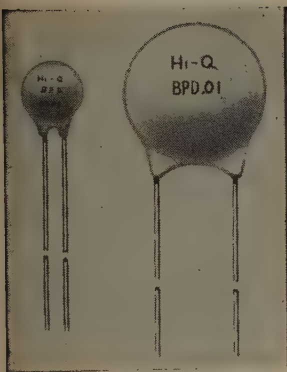

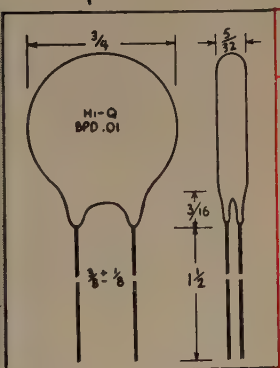

Type	A Diameter	B Lead Width	C Thickness
B.P.D. .00047	5/16" max.	3/16" ± 1/16"	5/32" max.
B.P.D. .0008	5/16" max.	3/16" ± 1/16"	5/32" max.
B.P.D. .001	3/8" max.	1/4" ± 1/16"	5/32" max.
B.P.D. .0015	3/8" max.	1/4" ± 1/16"	5/32" max.
B.P.D. .002	7/16" max.	1/4" + 1/8" − 0	5/32" max.
B.P.D. .004	19/32" max.	1/4" + 1/8" − 0	5/32" max.
B.P.D. .005	19/32" max.	1/4" + 1/8" − 0	5/32" max.
B.P.D. .01	3/4" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 2x.001	19/32" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 2x.0015	19/32" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 2x.002	19/32" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 2x.003	3/4" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 2x.004	3/4" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 3x.0015	3/4" max.	3/8" ± 1/8"	5/32" max.
B.P.D. 3x.002	3/4" max.	3/8" ± 1/8"	5/32" max.

Insulation: Durez and Wax impregnated.
Leads: 22 gauge pure tinned dead soft copper.
Capacity: Guaranteed mininum as stamped.
All capacitance measurements made at 25°C at 1 KC at a test voltage not over 5 volts RMS.

Insulation Resistance: 7500 megohms min.
Power Factor: Max. 2.5% at 1 KC at not over 5 volts RMS.
Test Voltage: 1500 volts D. C.

HI-Q COMPONENTS
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SALES OFFICES: New York, Philadelphia
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Jessup, Pa., Myrtle Beach, S. C.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

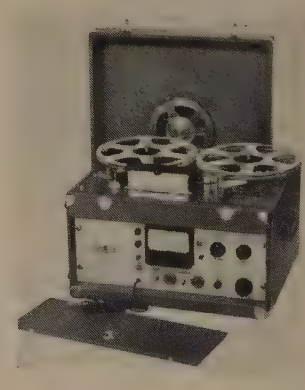
Aircraft Signal Generator

A signal generator for complete testing of airborne omnirange and localizer receivers in aircraft or on the bench has been introduced by the **Aircraft Radio Corp.,** Boonton, N. J.

The unit, designated as Type H-14, has a frequency range of 108 to 118 Mc, and can be used for all necessary quantitative bench tests. It provides facilities for testing 24 omni courses, plus left-center-right checks on both amplitude and phase localizers. The output for outside checks is 1 volt into 52 ohm line, and for bench checks, 0 to 10,000 μv.

New Tape Recorder

A high-fidelity magnetic tape recorder, with a range of 15,000 cps on half-track tape recorded at $7\frac{1}{2}$ inches per second is in production at **Ampex Electric Corp.,** 131 Howard Ave., San Carlos, Calif.



The capacity of this Model 400 is 132 minutes on a single 10-inch reel.

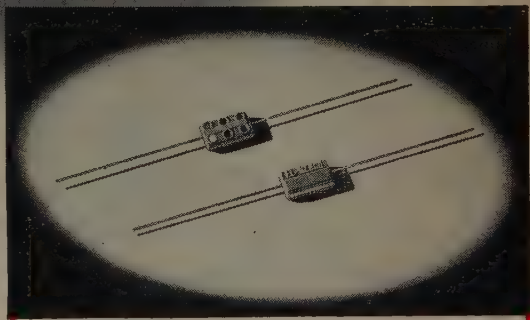
Other features include three magnetic heads shielded in a single housing, a built-in vu meter, and a single control switch for forward, rewind, and record. The machine also provides 15 inches per second tape speed by simply turning a switch. At $7\frac{1}{2}$ inches per second flutter and wow is less than 0.25 per cent; frequency response is ± 4 db, 30 to 15,000 cps. At 15 inches per second flutter and wow is less than 0.2 per cent; frequency response is ± 2 db, 50 to 15,000 cps.

(*Continued on page 22A*)

Only A SMALL PIECE OF *the pie* . . .

In the price of the El-Menco CM-15 capacitor, the cost of materials is small — for few materials are used. It's the know-how of putting these minute quantities of materials together that really counts.

Tiny as it is, the El-Menco CM-15 high-capacity fixed mica condenser exceeds the strict requirements of the Army and Navy. It is tested for dielectric strength at double its working voltage *before* leaving the factory — for insulation resistance and capacity value. You can *always* depend on this mighty midget — even under the most critical operating conditions and climate extremes.



CM-15 MINIATURE CAPACITOR

Actual Size 9/32" x 1/2" x 3/16"
For Television, Radio and other Electronic Applications.
2 mmf. to 420 mmf. cap. at 500v DCw.
2 mmf. to 525 mmf. cap. at 300v DCw.
Temp. Co-efficient ± 50 parts per million per degree C for most capacity values.
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ARCO ELECTRONICS, INC. 103 Lafayette St., New York, N. Y.—Sole Agent for Jobbers and Distributors in U.S. and Canada

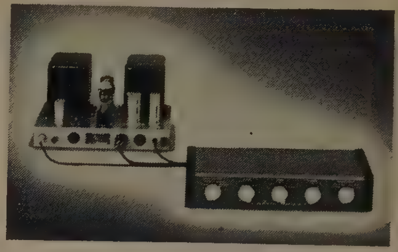
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 18A)

New Amplifier

A new wide-range, low-distortion amplifier system A-20-5 is available from **The Electronic Workshop,** 351 Bleecker St., New York 14, N. Y.



The four input channels, including an equalizer-preamplifier for any of the available magnetic phono cartridges, have independent level adjustments. A four-position treble cut-off filter reduces high-frequency noise and distortion. A loudness-compensated volume control is provided, as well as separate bass and treble controls giving 18-db boost or cut. The treble control equalizes for high-frequency recording characteristics.

A noise suppressor can be conveniently connected into the system to be effective on all channels. A recorder output which is unaffected by settings of the tone and volume controls is included.

The 18 db of feedback in the amplifier affords excellent loudspeaker damping and long tube life. Distortion at 20 watts is less than 1 per cent. Full power is delivered over the entire audio range.

A somewhat simplified system, the S-20 is also available.

Television Deflector Yoke Core

A high-permeability core for deflection of wide-angle, large-screen, kinescopes is available from **Westinghouse Electric Corp.,** P. O. Box 2099, Pittsburgh 30, Pa.

Made of Hipersil, a cold-rolled, grain-oriented electrical steel, the core has low reluctance at all flux densities.

The core is wound and bonded in circular form from a continuous strip of 5-mil material. It is then cut into two "C" shaped pieces for assembly around the deflection coils. The firm claims that the extremely thin laminations plus superior magnetic characteristics of the steel result in improved linearity and sharper pictures. The cores are rugged and completely free from magnetic instability due to change in temperature.

The core is available in sizes to suit the application.

(Continued on page 23A)

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 22A)

Plug-In Electronic Chassis

To meet a variety of requirements for compact, lightweight electronic assemblies, **Avion Instrument Corp.,** 121 E. 24 St., New York 10, N. Y. has designed a universal electronic chassis for use in servomechanisms, pulse and flip-flop circuits, analog computers, and similar devices. Up to eight subminiature tubes can be mounted and wired.

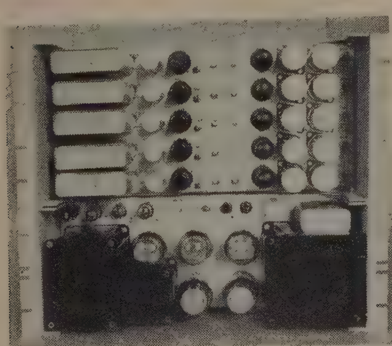
The assembly mounts in a metal case, filled with a potting compound, which provides mechanical support for the components. Input and output leads are brought to a standard octal tube base, for connection to associated equipment. Operating weight is 12 ounces.

The chassis is designed to operate in temperatures ranging from −50° C to +80° C. It meets specification AN-E-19 with respect to condition of altitude, humidity, and vibration.

Typical assembly is a direct-coupled basic amplifier having closed-loop characteristics as follows: output, with no load, −125 to 275 volts and, with 10^4 ohm load, −40 to 40 volts; output impedance, looking into amplifier, 1 ohm (approx.); linearity, variation over entire output range, less than 10 mv; gain is determined by choice of external resistors, maximum useful gain, limited by drift, 25, (approx.); drift, referred to input, 20 mv (approx.); frequency response 3 db down at 100 cps; operating voltages must be regulated to ±1 per cent.

Video Distribution Amplifier

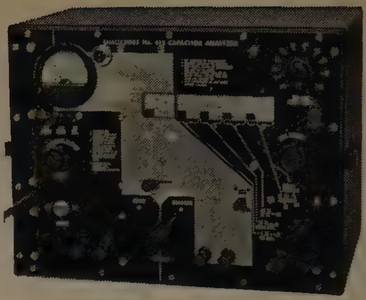
The Type 1311 video distribution amplifier which consists of five separate isolation amplifiers and an electronically regu-



lated power supply is in production at **Tel-Instrument Co., Inc.,** 50 Paterson Ave., East Rutherford, N. J.

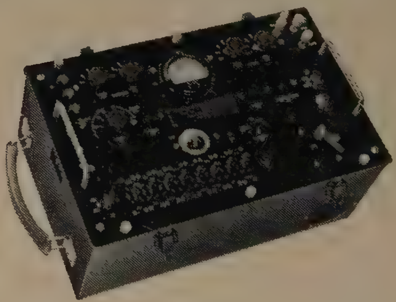
(Continued on page 74A)

Something New



WIDE-RANGE, DIRECT READING CAPACITOR ANALYZER

A laboratory-type Capacitor Analyzer meeting the need for a highly accurate, wide-range, direct-reading measuring instrument capable of determining the essential characteristics of capacitors has been announced by the Shallcross Manufacturing Co. This versatile instrument will determine capacitance values between 5mmf. and 12,000 mfd.; insulation resistance from 1.1 to 12,000 megohms; also leakage current, dielectric strength, and percentage power factor. A divided panel carrying an outline of the operating instructions makes it readily possible to use the instrument without reference to an instruction book. The Shallcross analyzer operates on 110 volt, 60-cycle alternating current. Literature giving full details will gladly be sent on request to the Shallcross Manufacturing Company, Collingdale, Pa.



MULTI-PURPOSE TRANSMISSION TEST SET

In addition to measuring the electrical characteristics of telephone lines and equipment the new Shallcross multi-purpose transmission test set may be used for efficiency tests on local and common battery telephone lines and sets, carbon microphones, receivers, and magnetic microphones. It also provides a fast, efficient means of testing capacitors, generators, ringers, insulation resistance, dials, and continuity. Key switches and dials are used to select and control the test circuits. The 693 Transmission Test Set is powered by external batteries. It features compact, substantial construction and is fully portable, thus making it ideal for either field or laboratory use. Details may be obtained from the Shallcross Manufacturing Company, Collingdale, Pennsylvania.

The DUMONT TYPE 297

Oscillograph-record Camera

You've just advanced the film after making an exposure. That starts the Polaroid-Land "packaged" developing process. And while you're waiting for its completion—in just 60 seconds—you can start the next exposure or set of exposures. With the Type 297, once you've snapped the shutter, its self-contained, 60-second, developing process lets you forget about the variables of the darkroom.

And with such special oscillographic features as simultaneous viewing and recording, an illuminated data-card, sturdy and easily attached mounting, and overall economy of the Polaroid film, Du Mont has added even greater meaning to this excitingly fast method for "printing" the oscillograph image.

SPECIFICATIONS

LENS—Du Mont-Wollensak f/2.8 or f/1.9, 75mm, coated.

SHUTTER — Wollensak Alphax; shutter speeds of 1/25, 1/50, 1/100 sec. Time and Bulb.

FOCUS—Fixed. May be adjusted for special oscillographic work.

WRITING SPEED—Writing rates of 3.5 in μsec. have been recorded consistently at 12,000 volts accelerating potential.

PRINT SIZE—3¼ x 4 ¼ in.—one, two, three, or more exposures per print.

IMAGE REDUCTION RATIO—2.25:1.

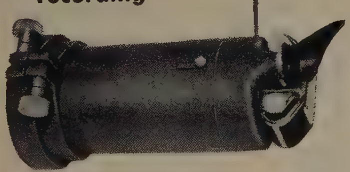
PHYSICAL SIZE—Length, 14⅝ in.; height, 10 in.; width, 6 in.

WEIGHT—12 lbs.

PRICE . . . . $285.00 with f/2.8 lens
$355.00 with f/1.9 lens

Write for bulletin on photographic techniques.

. . . and for thrifty SINGLE-FRAME recording



THE DU MONT TYPE 296

Oscillograph-record Camera

The Type 296 is an inexpensive oscillograph-record camera, greatly improved for general-purpose application with any standard 5-inch cathode-ray oscillograph. It incorporates a compact, all-metal, 35mm camera, calibrated shutter and a high-quality f/2.8, 75mm, coated lens which increases its capability 57% over the Type 271-A which it supersedes. Construction is rugged and durable; operation simple and foolproof. The Type 296 weighs only 4½ lbs.

PRICE . . . $149.50

ALLEN B. DU MONT LABORATORIES, INC., *INSTRUMENT DIVISION*, 1000 MAIN AVENUE, CLIFTON, NEW JERSEY

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 23A)

Recent Catalog

• • • **Parts Section, Receiver Div., General Electric Co.,** Electronics Park, Syracuse, N. Y., has prepared a phono accessory catalog with descriptions and data on variable reluctance cartridges, replacement baton styli, tone arms, and phono preamplifiers.

Electronic Counter Tachometer

The **Potter Instrument Co., Inc.,** 115 Cutter Mill Rd., Great Neck, L. I., N. Y. announces a new high-speed electronic counter tachometer for high accuracy revolution and frequency measurements.

Fundamentally, the instrument totalizes the number of counts derived from the source being measured during a precisely established time interval of 0.6 second (0.01 minute). This reading is displayed on a direct-reading, four-digit electronic counter using ten neon glow lamps for each digit. After each measurement and display period, the instrument automatically resets and recycles. The display time is adjustable over a period of from 0.5 second to 4 seconds, or can be set to hold the count indefinitely.

The 0.6-second time interval is established by counting 60,000 cps of the 100-kc crystal oscillator contained in the unit.

The instrument can be used to measure frequencies up to 100,000 cps with an accuracy of ±1 cps. In addition to the electronic counter registration of the measured frequency, the unit also includes a count rate meter which can be used when the unknown frequency source is being adjusted. This type of indication reflects instantaneous changes and is useful when the cource of frequency must be manually adjusted to a specific point.

Two-Signal Audio Generator

A new Type 1303-A, two-signal audio generator, designed specifically for supplying the test signals necessary in the various methods of measuring intermodulation distortion in audio systems, has been developed by **General Radio Co.,** 275 Massachusetts Ave., Cambridge 39, Mass.

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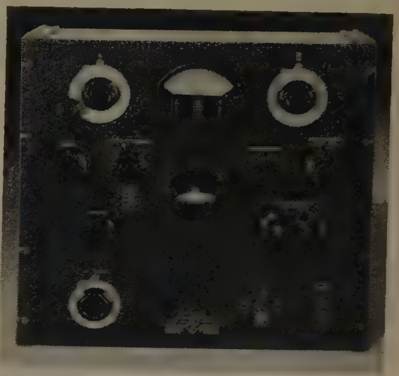
This generator can also be used as a general purpose laboratory beat-frequency oscillator. This oscillator will supply any of the following signals: A single low-distortion sinusoidal voltage adjustable in frequency from 20 cps to 40 kc in two ranges, 20 cps to 20 kc, and 20 kc to 40 kc; two low-distortion sinusoidal voltages,



each separately adjustable, one to 20 kc and the other to 10 kc and two low-distortion sinusoidal voltages, with a fixed difference in frequency maintained between the two as the frequency of one voltage is varied. The fixed difference frequency is adjustable up to 10 kc, and the lower of the two frequencies is adjustable up to 20 kc.

The output of the oscillator is continuously adjustable up to 10 milliwatts into 600 ohms with less than 0.25 per cent distortion, and up to 1 watt with less than 0.5 per cent distortion. Output is calibrated both in volts and in db with respect to 1 milliwatt in 600 ohms.

Plant Expansion

Radio Receptor Co., Inc., 84 N. 9th St., Brooklyn 11, N. Y., announces purchase of a 90,000 square foot factory at Wythe Ave. & N. 3 St., Brooklyn, N. Y., to increase space for their manufacture of rectifiers, hf heat sealing machines, and airport radio equipment.

Nuclear Test Equipment

Four new radiation detecting instruments, are now available from the **General Electric Co.,** Electronics Park, Syracuse, N. Y.

The small radiation monitor measures the total amount of gamma radiation in a given area over a period of time. It is self-contained, and because it has no batteries or tubes is not susceptible to moisture or dirt. Charging is immediately accomplished by inverting the device.

The portable alpha survey meter completely self-contained, is for the detection and indication of alpha radiation. It utilizes a proportional counting chamber as the detector. Output is given quantitatively on a meter and audibly on head-

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News—New Products

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phones. The semilog meter covers two decades without range switch. Power supply is by storage type batteries. The instrument is calibrated from four to five mev.

The explosion proof ionization chamber instrument is a survey meter suitable for use in class 1, group D hazardous atmospheres. Indication of all types of radiation is on a standard 3½ inch instrument. It is pressurized to 15 pounds per square inch and has a pressure relief valve and pressure operated switches to open battery circuit upon loss of pressure. Conductive rubber bumpers insure against accumulation of electrostatic charge.

The beta-gamma survey meter utilizes two Geiger-Mueller tubes operating from an electronic high-voltage supply to provide high sensitivity over a wide range of levels. The compact unit features a scale changing meter, the selecting switch also selecting the G-M tube to be used for each of the four ranges from 0 to 500 milliroentgens per hour. A detachable probe also contains a G-M counter. Battery voltage is regulated electronically to provide long-term metering accuracy, and battery condition is indicated on a meter. The instrument is furnished calibrated in milliroentgens per hour using a radium source. Calibration of each range can be individually varied.

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News—New Products

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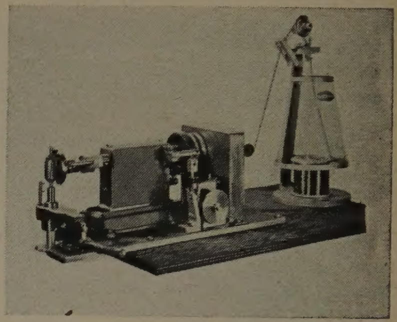
New Small Armature Winder

A new Model 36 armature winder, designed to wind small armatures and featuring an exclusive stationary armature, has just been announced by **Geo. Stevens**



Mfg. Co., Inc., 6022 N. Rogers Ave., Chicago 30, Ill. Wire is fed through a hollow spindle to a revolving arm which winds wire in the armature slots, permitting armature to remain stationary so that it cannot fly out and injure the operator.

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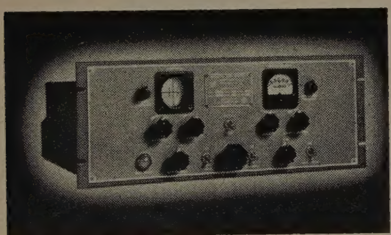
The machine operates at high speed, winding 1,000 turns per minute on either straight or skewed armatures. Very tight armatures may be wound because of the uniform tension with which the wire is automatically guided and laid in the slots. Amount of tension is limited solely by the strength of the wire. Positive control over number of turns is attained by automatically winding a predetermined number of turns. Armature is also automatically indexed.

Frequency Shift Converter

A small frequency shift converter has been designed and manufactured by **Northern Radio Co., Inc.,** 143 W. 22 St., New York 11, N. Y., for single and diversity receiving systems.



The Type 107 Model 2 is a dual channel unit which converts mark and space tones into dc pulses and drives teleprinters, tape and other recorders directly. Its integral 2-inch oscilloscope provides a tuning pattern for precise receiver adjustment, during initial setup and while keying.

This unit may also be used as a make and break cw or icw demodulator.

Other specifications of this unit are: input FS limits, 100- to 1,000-cps frequency shift; mark frequency set at 2,975 cps. The cw-icw input, 400 to 7,500 cps. Output (1) neutral dc pulses, 60 ma in 1,800 ohm load, one side grounded; (2) polar dc pulses, ±30 ma in 1,800 load, center grounded; (3) may be operated into any impedance from 100 to 100,000 ohms.

Compact Four-Decade Audio Oscillator

The new Model 510-A audio oscillator, with a frequency range from 18 to 210,000 cps in four decades, is available from **The Electronic Workshop, Inc.,** 351 Bleecker St., New York 14, N. Y.

The unit (dimensions $4 \times 5\frac{1}{2} \times 4$ inches) will deliver 10 volts into 10,000 ohms, with output constant within 0.5 db over the entire frequency range. Distortion at this amplitude is less than 0.3 per cent from 100 to 15,000 cps and rises to no more than 0.5 per cent at 30 cps. Source impedance of the cathode-follower output is 560 ohms; a matching transformer will be made available to feed low-impedance balanced lines. The total frequency error due to drift and dial calibration is less than ±2 per cent; the 320° dial gives a scale length

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News—New Products

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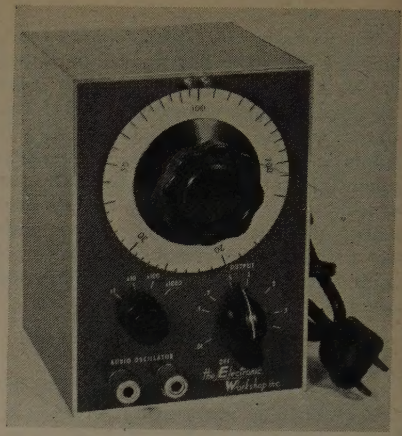

of over eight inches. The output control is logarithmic and is calibrated approximately in output voltage.

Recent Catalogs

• • • **Allied Electric Products, Inc.,** and **Sheldon Electric Co., Division,** 68–98 Coit St., Irvington 11, N. J., have printed a comprehensive 1950–1951 catalog #161, with technical data and prices concerning electrical cords, plugs, female outlets, television tubes, and lamps.